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FOR DESIGNERS AT HIGHER FREQUENCIES

Microwaves & RF

Semiconductor Issue

NEWS
Semiconductors vie for space in wireless systems

DESIGN FEATURE
A primer on using PIN diodes in VCAs

PRODUCT TECHNOLOGY
InGaP HBTs promise long operating lifetimes

Broadband synthesizer trims phase noise through 40 GHz

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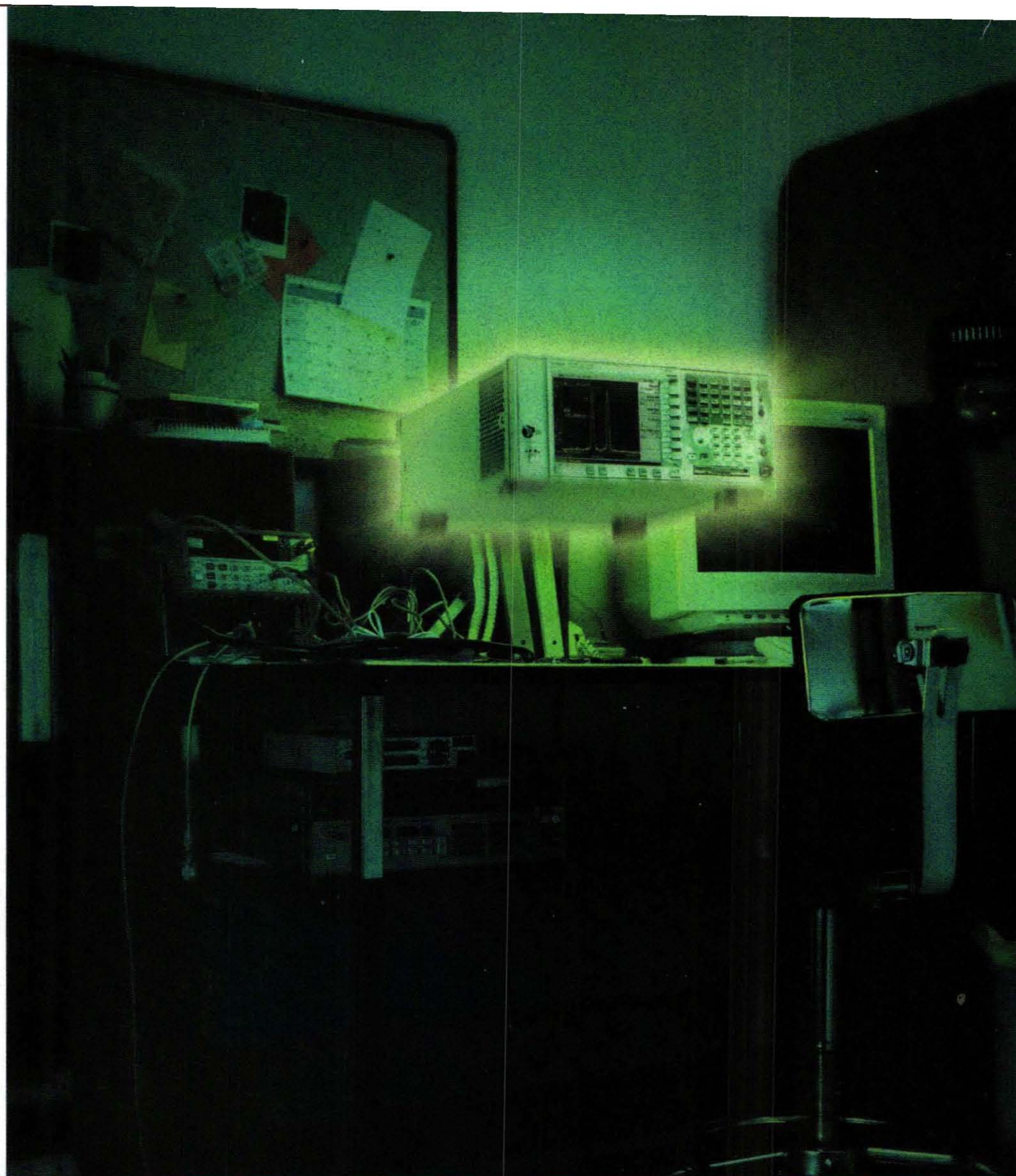
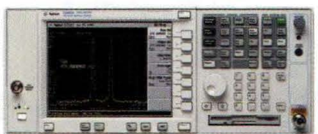
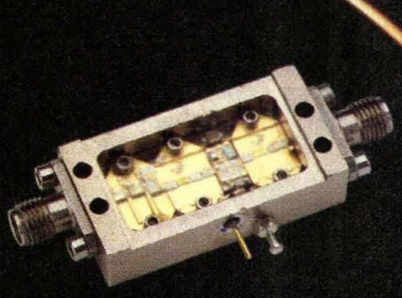
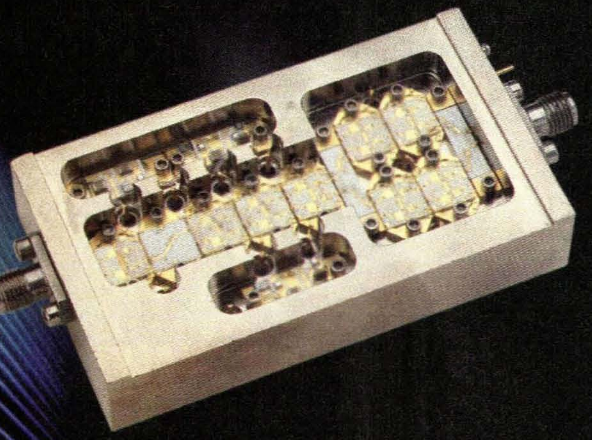
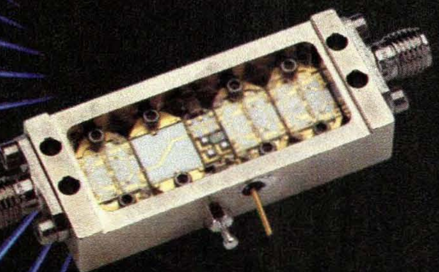




AMPLIFIERS for every application







ULTRA BROAD BAND

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA018-203	0.5-18.0	20	5.0	2.5	7	17	2.0:1	250
JCA018-204	0.5-18.0	25	4.0	2.5	10	20	2.0:1	300
JCA218-506	2.0-18.0	35	5.0	2.5	15	25	2.0:1	400
JCA218-507	2.0-18.0	35	5.0	2.5	18	28	2.0:1	450
JCA218-407	2.0-18.0	30	5.0	2.5	21	31	2.0:1	500

MULTI OCTAVE AMPLIFIERS

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA04-403	0.5-4.0	27	5.0	1.5	17	27	2.0:1	550
JCA08-417	0.5-8.0	32	4.5	1.5	17	27	2.0:1	550
JCA28-305	2.0-8.0	22	5.0	1.0	20	30	2.0:1	550
JCA212-603	2.0-12.0	32	5.0	3.0	14	24	2.0:1	550
JCA618-406	6.0-18.0	20	6.0	2.0	25	35	2.0:1	600
JCA618-507	6.0-18.0	25	6.0	2.0	27	37	2.0:1	800

MEDIUM POWER AMPLIFIERS

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-P01	1.35-1.85	35	4.0	1.0	33	41	2.0:1	1000
JCA34-P02	3.1-3.5	40	4.5	1.0	37	45	2.0:1	2200
JCA56-P01	5.9-6.4	30	5.0	1.0	34	42	2.0:1	1200
JCA812-P03	8.0-12.0	40	5.0	1.5	33	40	2.0:1	1700
JCA1218-P02	12.0-18.0	22	4.0	2.0	25	35	2.0:1	700

LOW NOISE OCTAVE BAND LNA'S

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-3001	1.0-2.0	40	0.8	1.0	10	20	2.0:1	200
JCA24-3001	2.0-4.0	32	1.2	1.0	10	20	2.0:1	200
JCA48-3001	4.0-8.0	40	1.3	1.0	10	20	2.0:1	200
JCA812-3001	8.0-12.0	32	1.8	1.0	10	20	2.0:1	200
JCA1218-800	12.0-18.0	45	2.0	1.0	10	20	2.0:1	250

NARROW BAND LNA'S

Model	Freq. Range GHz	Gain dB min	N/F dB max	Gain Flat +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-1000	1.2-1.6	25	0.75	0.5	10	20	2.0:1	80
JCA23-302	2.2-2.3	30	0.8	0.5	10	20	2.0:1	80
JCA34-301	3.7-4.2	30	1.0	0.5	10	20	2.0:1	90
JCA56-401	5.4-5.9	40	1.0	0.5	10	20	2.0:1	120
JCA78-300	7.25-7.75	27	1.2	0.5	13	23	2.0:1	120
JCA910-3000	9.0-9.5	25	1.2	0.5	13	23	1.5:1	150
JCA910-3001	9.5-10.0	25	1.2	0.5	13	23	1.5:1	150
JCA1112-3000	11.7-12.2	27	1.1	0.5	13	23	1.5:1	150
JCA1213-3001	12.2-12.7	25	1.1	0.5	10	20	2.0:1	200
JCA1415-3001	14.4-15.4	35	1.4	1.0	14	24	2.0:1	200
JCA1819-3001	18.1-18.6	25	1.8	0.5	10	20	2.0:1	200
JCA2021-3001	20.2-21.2	25	2.0	0.5	10	20	2.0:1	200

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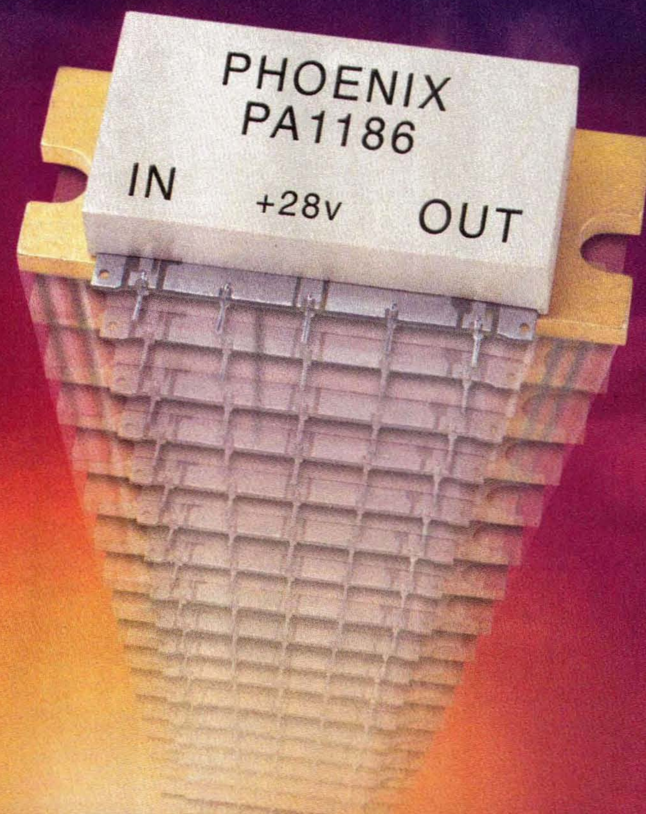

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Parameter	PA1186	PA1166	PA8106	PA8105	PA1182
Frequency (MHz.)	800-960	1930-1990	2000-2200	2110-2170	2300-2400
Gain (dB.)	29.0	26.0	25.0	25.0	23.0
Pout @ 1 dB. comp. (dBm.)	38.0	38.0	37.5	37.5	37.5
Noise Figure (dB.)	2.4	2.7	3.0	3.0	3.0
ACPR (30kHz BW)*	-50.0	-54.0	-47.0	-47.0	-47.0
VSWR (Input/Output)	1.5:1/2:1	1.5:1/2:1	1.5:1/2:1	1.5:1/2:1	1.5:1/2:1
IP3 (two tone)**	+56.0	+54.0	51.0	51.0	51.0
Supply Required	+28/1000	+28/1000	+28/1000	+28/1000	+28/1000

* ±850kHz from fc at power level of 30 dBm. (IS-95)
** IP3 measured with 2 tones @ +25dBm. per tone

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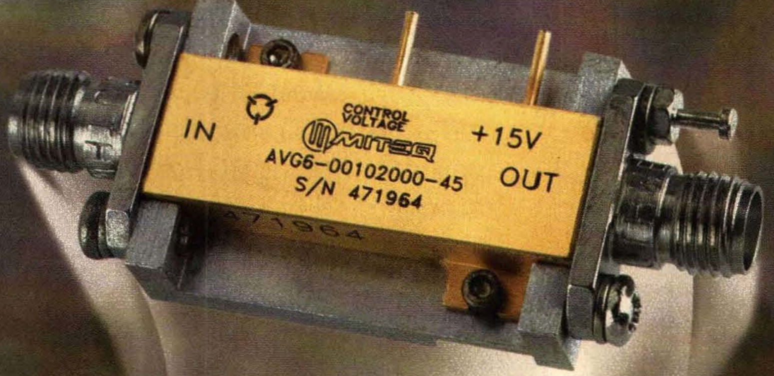
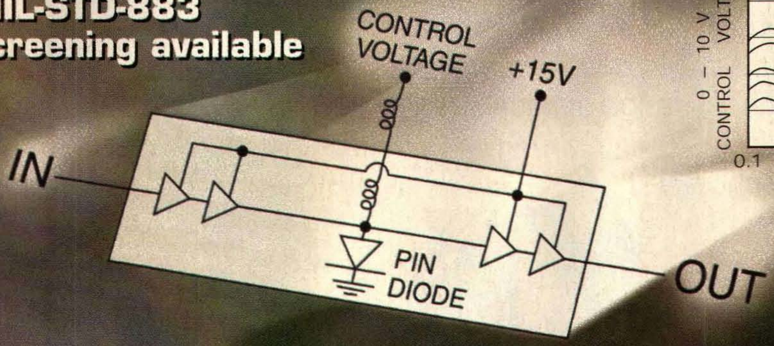
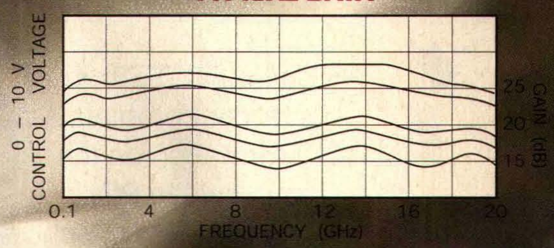




TYPICAL DATA



3 YEAR WARRANTY

MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN FLATNESS (dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN/OUT (Max.)	OUTPUT POWER @ 1 dB Comp. (dBm, Min.)	NOM. DC POWER (+15 V, mA)
AVG4-00100400-14	.1–4	28	±1.00	1.4	2.0:1	+10	150
AVG4-00100600-15	.1–6	28	±1.00	1.5	2.0:1	+10	150
AVG4-00100800-18	.1–8	26	±1.50	1.8	2.0:1	+10	175
AVG4-02000800-20	2–8	32	±1.25	2.0	2.0:1	+10	175
AVG5-04000800-12	4–8	30	±1.00	1.2	2.0:1	+10	150
AVG5-00101800-35	.1–18	24	±2.50	3.5*	2.5:1	+10	175
AVG6-00102000-45	.1–20	24	±2.50	4.5*	2.5:1	+10	250
AVG4-06001200-19	6–12	24	±1.50	1.9	2.0:1	+10	175
AVG4-06001800-25	6–18	22	±2.00	2.5	2.3:1	+10	185
AVG6-02001800-40	2–18	25	±2.25	4.0	2.5:1	+10	250

* Noise figure increases below 500 MHz.

Note: All above specifications are with 0 dB attenuation.

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REPORT CLARIFICATION

To the editor:

I enjoyed the Special Report on power devices by Barry Manz that appeared in the November issue (pp. 125-132). There were some points made in the article, however, that need to be clarified.

It was former House Speaker Newt Gringrich who first stood before a Senate committee, holding an octal-based receiving tube and not a miniature 12AU7 (*Science*, March 17, 1995). He was making his point with an example supposedly from air-traffic-control radar. The former Speaker missed the point as to why high-powered radar transmitters might continue to need tubes—in the power-amplifier (PA) stages. Manz should also check his data as the 12AU7 was a dual triode used in consumer and industrial applications as a low-frequency amplifier and oscillator, but not as a rectifier.

In another paragraph, Mr. Petrini of CPI says that the AM radio market is practically solid state to approximately 15 kW. All-transistor AM radio transmitters have been delivered to customers with power levels that reach up to 1 MW of carrier power. Plastic-packaged vertical-channel metal-oxide semiconductor field-effect transistors (MOSFETs) are typically used with high-efficiency circuit topologies to deliver amplitude-modulated RF to an antenna for systems operating below 2 MHz. Solid-state 50-kW transmitters are standard products for these medium wavelengths.

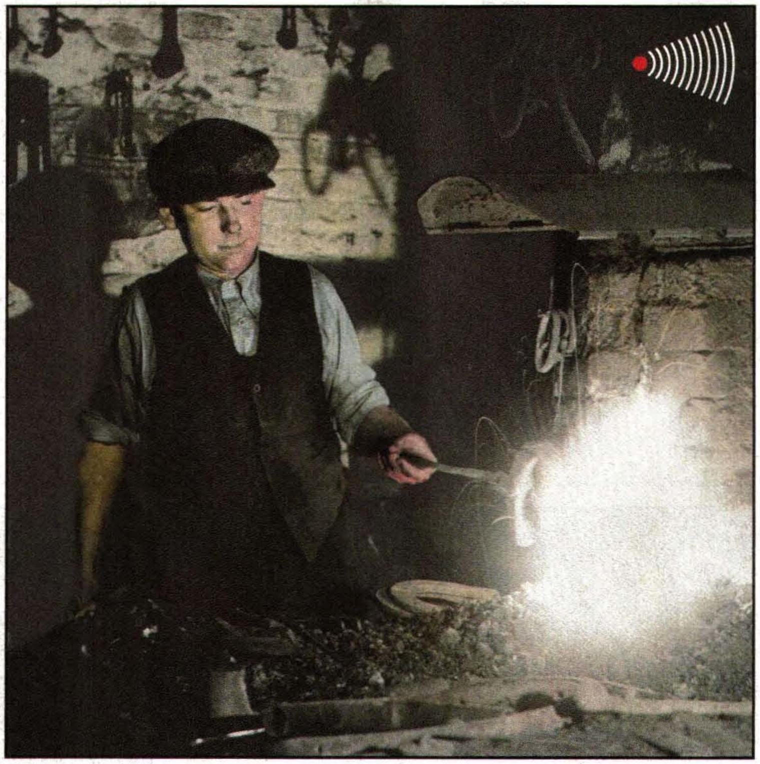
Laterally diffused metal-oxide semiconductor and vertical-channel DMOS technologies rule for the very-high-frequency (VHF) and ultra-high-frequency (UHF) ranges, but the solid-state devices are more expensive due to their controlled impedance, packaging, and mode of operation, which generates more dissipated power as heat. Single-tube FM transmitters continue to be available for 25-to-35-kW levels. Solid-state PAs are used as drivers for the tubes in these rigs, so they could be considered hybrids.

High-definition TV does not employ klystrons as Manz states, but uses the Inductive Output Tube (IOT) being manufactured from a number of sources. Klystrons have lost their place in the market as UHF analog (NTSC) television plants are being replaced.

In the paragraph about using klystrons as a source of RF power for irradiation of meat, it must be noted that the irradiation comes from a beam of ionized particles, such as electrons, being accelerated in a machine, which uses RF power. The report implied that the klystrons would irradiate the meat. RF power uses that are exclusively in the domain of high-power thermionic tubes are heavy particle accelerators used for dielectric heating. Klystrons and power-grid tubes are used in particle accelerators, most being larger machines than the medical linacs mentioned in the report.

John Lyles

FABRICATION



GALLIUM ARSENIDE

COULD THIS BE THE YEAR OF BLUETOOTH?

Bluetooth lovers were in their element last month. The Bluetooth Developers Conference (December 5-7, 2000, San Jose Convention Center, San Jose, CA) began as a little get-together of parties interested in this 2.4-GHz wireless personal connectivity technology, and finished as a genuine, full-blown conference with apparent satisfaction from engineering-oriented attendees and market-driven exhibitors. If the total attendance of this conference/exhibition is any measure (approximately 3000), then the future appears quite bright for Bluetooth.

The Bluetooth Developers Conference featured a nice mix of technology (in the meeting rooms) and sales hype (on the exhibit floor). The meeting was presented by Bluetooth Special Interest Group (SIG) members 3COM, Ericsson, IBM, Intel, Lucent Technologies, Microsoft, Motorola, Nokia, and Toshiba, and was produced and managed by Key3Media Events (Needham, MA).

The technology was evident at some well-prepared technical presentations. These were organized into three key areas: hardware, software, and test equipment. Since the hardware, in the form of integrated circuits (ICs) seems further along than either the software or the test solutions, these papers tended to be reports on existing and emerging products, with some measure of salesmanship added. Many of the papers on software and test addressed the need for interconnectivity in order for Bluetooth to earn widespread acceptance in consumer markets. The Bluetooth standard, which details hardware, software, and test requirements, is fairly firm in its published specifications for hardware requirements. This has allowed the development of a variety of IC solutions for both Bluetooth radios and the baseband controllers.

On the show floor, the energy level was high. Fueled by some wild market research studies that predicted global demands for one billion and more Bluetooth units by 2005 [does this seem vaguely like the market projections in the early 1980s for gallium-arsenide (GaAs) devices?], IC suppliers were enthusiastic about the coming volumes. Engineers at the Infineon Technologies (Munich, Germany) booth, for example, were quoting customers prices for one million-piece orders. Ironically, across the hall at Crossbow Technology (San Jose, CA), a supplier of Bluetooth-based sensors for process monitoring, engineers were bemoaning the slow trickle of Bluetooth modules from their supplier.

Although there was some skepticism about the potential of Bluetooth markets, on the exhibit floor and in the technical sessions, the belief in the future of Bluetooth was strong at this conference. Given the technical and marketing strengths of the Bluetooth SIG, and the 2000 member companies (both small and large), it is hard to imagine that Bluetooth does not become a major market. For more on Bluetooth, watch for next month's Special Report (wrapup) on the Bluetooth Developers Conference.

Jack Browne

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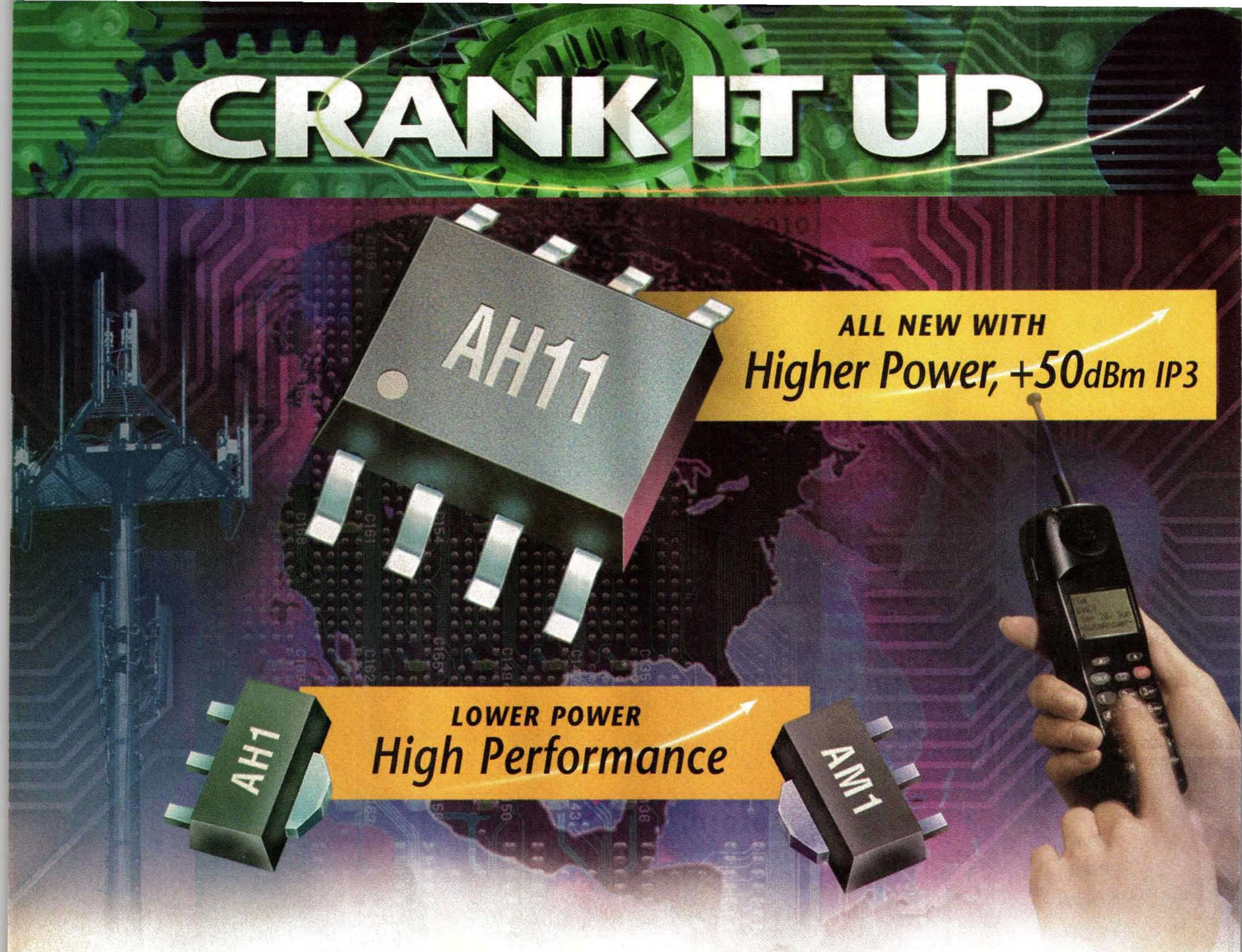

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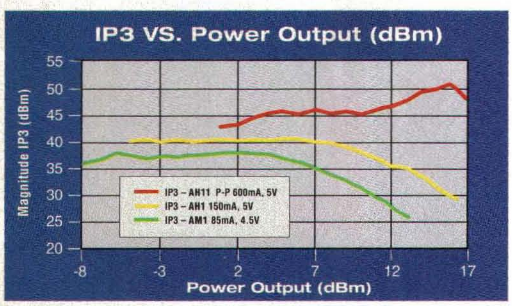

WJ High Dynamic Range Amplifiers

Product	Frequency (MHz)	IP3 (dBm, typ.)	P1dB (dBm, typ.)	NF (dB, typ.)	Bias current (mA, typ.)
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AH1	250-3000	41	21	2.9	150
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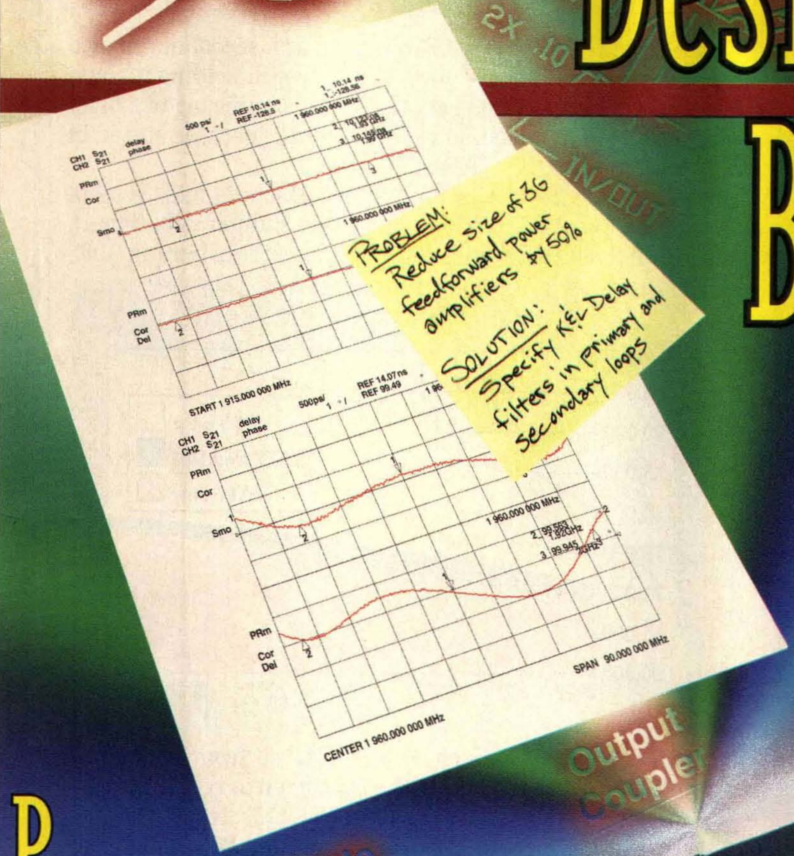
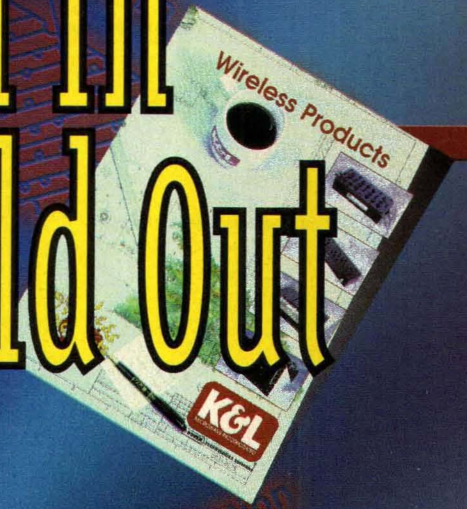
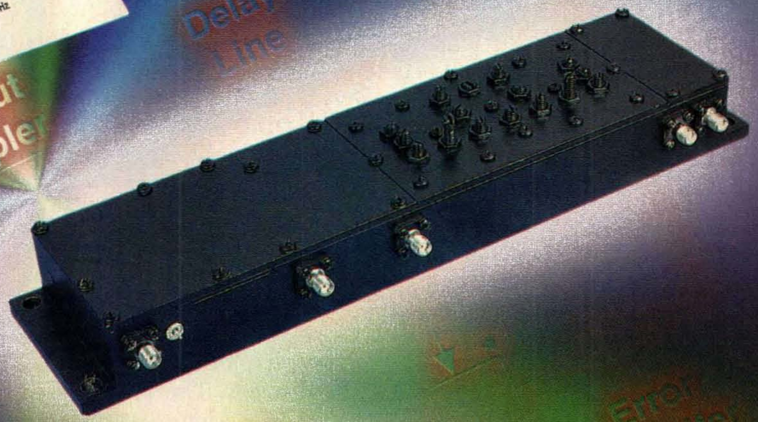
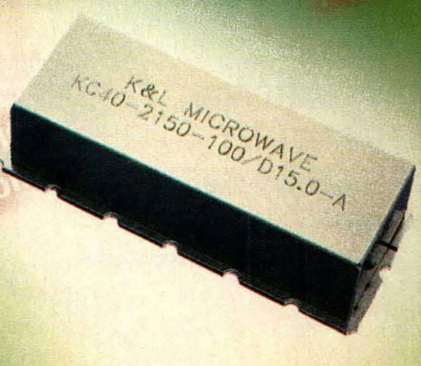






Primary loop delay filter assemblies integrate high power cavity delay filters, couplers and isolators into a single compact module for performance improvement, space savings and cost reduction.



Secondary loop miniature delay filters utilize a Patent Pending topology to provide delay equalization. The filters exhibit excellent delay and phase flatness in a miniature surface mount package. The 3G model measures a mere 1.5"L x 0.60"W x 0.4"H and can be mounted directly onto your PWB with pick and place machinery.

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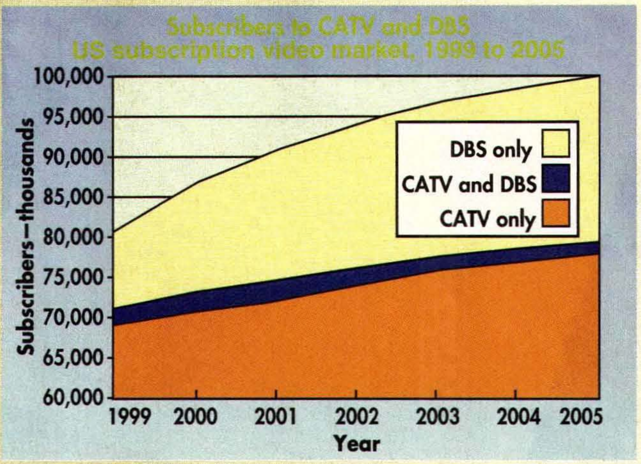
CATV Will Meet DBS Challenge

OYSTER BAY, NY—While cable television (CATV) is seen as a secondary choice to direct-broadcast satellite (DBS) for new subscribers, CATV is still holding its own despite dire predictions. However, the year 2000 represented a significant but brief shift as new DBS subscribers more than doubled new CATV subscribers, as stated in a report from Allied Business Intelligence, Inc. (ABI).

According to "CATV Infrastructure 2000: US Equipment Markets and System Trends," an annual research study by ABI, CATV added two million subscribers in 1999 and another two million in 2000. DBS added 2.7 million subscribers in 1999 and 4.5 million in 2000 (see figure).



However, DBS subscribership will never be on a 50-50 basis with cable. Even on a new subscriber basis, cable will exceed DBS from 2002 onward as the initial effect of carrying local channels is spent.

Despite the DBS segment gain, the DBS industry is still new enough to have a relatively low subscriber count—just over 11 million at the end of 1999, versus CATV's 71 million. It has suffered from the inability to offer local channels to the customer. This forced the customer to rely on broadcast antennas or even the cable company to obtain those local channels. Nearly 1.6 million households subscribed to DBS and CATV in 1999, according to ABI's findings.

This changed in 2000. Recent legislation and technological advances will allow DBS operators to offer local channels in many major markets.

While this will increase DBS's attractiveness, it does not remove all of cable's advantages. For one, cable already has a large customer base. These customers must make an effort to switch to DBS, while no effort is necessary to keep the existing cable services. The existing subscriber base also provides cable operators with data bases consisting of names and potential revenues for marketing.

Researchers Work To Enhance Computer Data-Storage Capabilities

AMES, IA—Research is currently being performed that will lead to the creation of a personal computer (PC) that stores 10 to 50 times more information than today's top models, does not lose power during power interruptions, and starts up immediately without needing the traditional "boot-up" process.

Breakthroughs such as these are being explored in the rapidly growing field of magnoelectronics, which combines microelectronics and magnetics to create new technologies that will quench the public's growing thirst for greater data-storage capacities on computers.

Scientists at the US Department of Energy's Ames Laboratory and Iowa State University will become part of those research efforts, thanks to a $530,000 grant from the Roy J. Carver Charitable Trust to establish a magnoelectronics laboratory.

The grant was awarded in June 2000 to David Giles and John Snyder, who are reseachers at Ames Laboratory and ISU. They are reviewing bids on equipment for the new lab, which will be located in Ames Laboratory's Metals Development building on the north side of ISU's campus.

"Magnetoelectronics is a very, very hot area. We've all got computers and we all want to be able to store more and more data on them," Jiles says. "Ames Laboratory and Iowa State need to get into this area because there's a huge market for this type of cutting-edge technology."

Much of the work in the new lab will involve using an ion-beam deposition system to produce materials in the form of thin films for technologies that will expand computer data-storage capacities.

Progress Made In Miniaturizing Fuel-Cell Power Sources

TEMPE, AZ—Scientists at Motorola Labs have reached another milestone in their development of a new miniature fuel cell that may one day replace the traditional batteries that now power everything from cellular phones and laptop computers to portable cameras and electronic games. They have demonstrated a prototype of a ceramic-based microfluidic fuel-delivery system for a miniature direct-methanol fuel cell (DMFC).

"Portable electronics are becoming more essential to daily life and, increasingly, we all want them to have new capabilities," says Jerry Hallmark, manager of Motorola Labs' Energy Technology Lab. "But adding features increases the demand on energy sources and systems. We need to develop new energy solutions—and fuel cells could be the breakthrough technology. Our challenge is to make these systems small, light, and easy for consumers to use. Eventually, these fuel cells could enable what people just dream of today—a lightweight energy source that would safely power a cellular phone for a month."

To produce energy, the new fuel cell uses a reservoir of inexpensive methanol that, when combined with the oxygen in the air, produces electricity at room temperature. Motorola's initial strategy is to develop a hybrid energy source, which combines a miniature fuel cell with a rechargeable battery for peak power demands.

Information-Based Economy Is Leveling The Technology-Exports Playing Field

ATLANTA, GA—A study of international competitiveness may provide US producers of technology products with another reason to be looking over their shoulders.

Though the US remains the undisputed world leader in exporting technology products, the Georgia Institute of Technology study of technological capabilities among 33 nations shows the industrializing countries of Asia quickly catching up, thanks to an information-based economy that facilitates rapid change. The National Science Foundation-sponsored study, "Indicators of Technology-Based Competitiveness," is the latest in a series of reports published every three years since 1987.

"Our study points to a much more competitive environment for the United States," says Dr. Alan Porter, director of the Technology Policy and Assessment Center at Georgia Tech. "The playing field is changing from a ski slope to a gentle plateau. No longer is the United States alone on the playing field with the Japanese."

Though the study evaluates nations as varied as Israel, Brazil, and the Czech Republic, Porter sees the real action among the "Asian Cubs." These challengers—including China, India, Malaysia, Thailand, Indonesia, and The Philippines—are moving up alongside traditional regional leaders, the "Asian Tigers" South Korea, Taiwan, and Singapore.

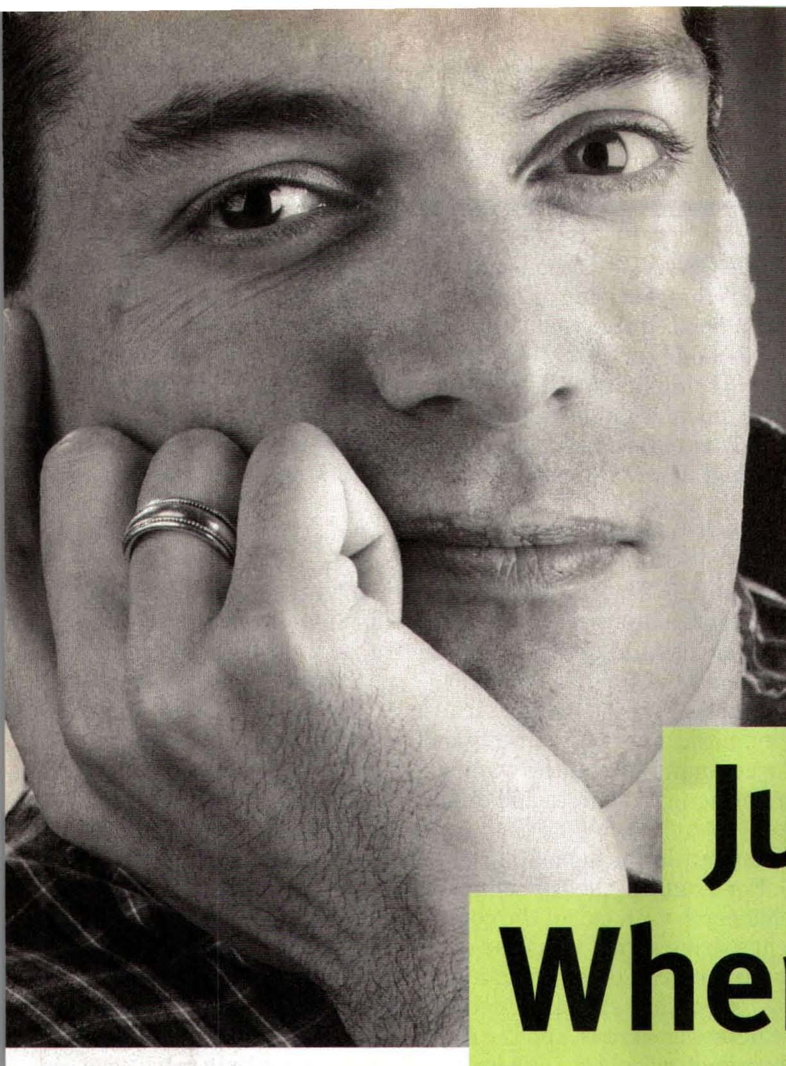
Over the past decade, these nations have developed the technological infrastructure to move from manufacturers of products to developers of products. The growth of indigenous engineering and management capabilities, development of research-and-development (R&D) capabilities, and the rise of entrepreneurship signal this transition.

One Billion Cellular Handsets To Be Shipped Yearly By 2004

WELLINGBOROUGH, NORTHANTS, ENGLAND—According to a market report by the wireless-communication-market specialist Intex Management Services Ltd. (IMS), the number of cellular terminals (handsets) shipped in a year is forecast to surpass the 1 billion mark in 2004. This is compared to an estimated market of 288 million terminal shipments in 1999.

Growth in the market is projected at 1.34 billion terminal shipments in 2006. At this time, it is forecast that there will be 1.79 billion users of cellular services worldwide. "By 2006, increases in users in developed countries is forecast to have started to peak. However, take up of cellular in less-developed countries, such as China, and regular handset replacements by users worldwide will mean that the market for cellular terminals will continue to experience significant growth," states Alex Green, a senior analyst for IMS and one of the report's authors.

Over the period from 1999 to 2006 there will be a marked change in the makeup of the types of cellular handsets that are sold. It is estimated that in 1999, approximately 95 percent of terminals shipped complied to second-generation (2G) standards with the rest being analog. However, it is forecast that 56 percent of handsets shipped will comply with 2.5G or third-generation (3G) standards.



Patent Awarded For Microsatellite Technology

PRINCETON JUNCTION, NJ—Discovery Semiconductors, Inc. announced the grant of US Patent 6,137,171 entitled "Lightweight miniaturized integrated microsatellite employing advanced semiconductor processing and packaging technology." The innovations claimed in the patent enable reducing the weight of communications and remote-sensing satellites to 22 lbs. (10 kg) and the volume to 305 cubic in. (5000 cubic cm). The instrumentation module or payload portion of the satellite can be reduced to 1.07 oz. and 12 cubic in. (32 g and 200 cc). This drastically lowers the cost of launching and maintaining fleets of low-earth-orbiting (LEO) satellites. The packaging concepts used within the patent have broad applications for reducing the size and weight of many complex electronic assemblies in terrestrial and submarine systems.

Conventional satellites are 10 times larger and heavier than the patented design, due to the bulk of packaging separate functions in individual subassembles. By employing the company's capabilities to design and fabricate optoelectronic integrated circuits (OE ICs) on gallium-arsenide (GaAs) and silicon (Si) wafers, complete subsystems of the electronic module will be fabricated on individual wafers. These wafers are then stacked in a cylindrical central housing and interconnected to the module through contacts on the circumference of the wafers. Selected signals can be communicated between wafers using light sources and detectors integrated onto the wafers through methods patented by Discovery Semiconductors in US patent number 5,621,227 ("Method and apparatus for monolithic optoelectronic integrated circuit using selective epitaxy"). For example, the RF communications antenna would be fabricated on the wafer facing earth with one of the company's wide-bandwidth detectors at its center. The antenna drive signal would be communicated from the signal-processing wafer through an integrated semiconductor laser, eliminating the need for bulky RF interconnects. In another example cited in the patent, the diffraction grating, several photodiode detection arrays, and the signal-processing circuitry of a multiwavelength-band spectrometer could be implemented on one wafer for remote-sensing applications. Combined with additional optics, the spectrometer would image the earth's landmass in ultraviolet (UV), visible, and infrared (IR) light to monitor the health of crops. The monolithic design greatly improves the reliability of the spectrometer as it eliminates many mechanical connections that would be susceptible to vibration failure in a space launch.

Platform Extends Indoor Wireless Communications Coverage

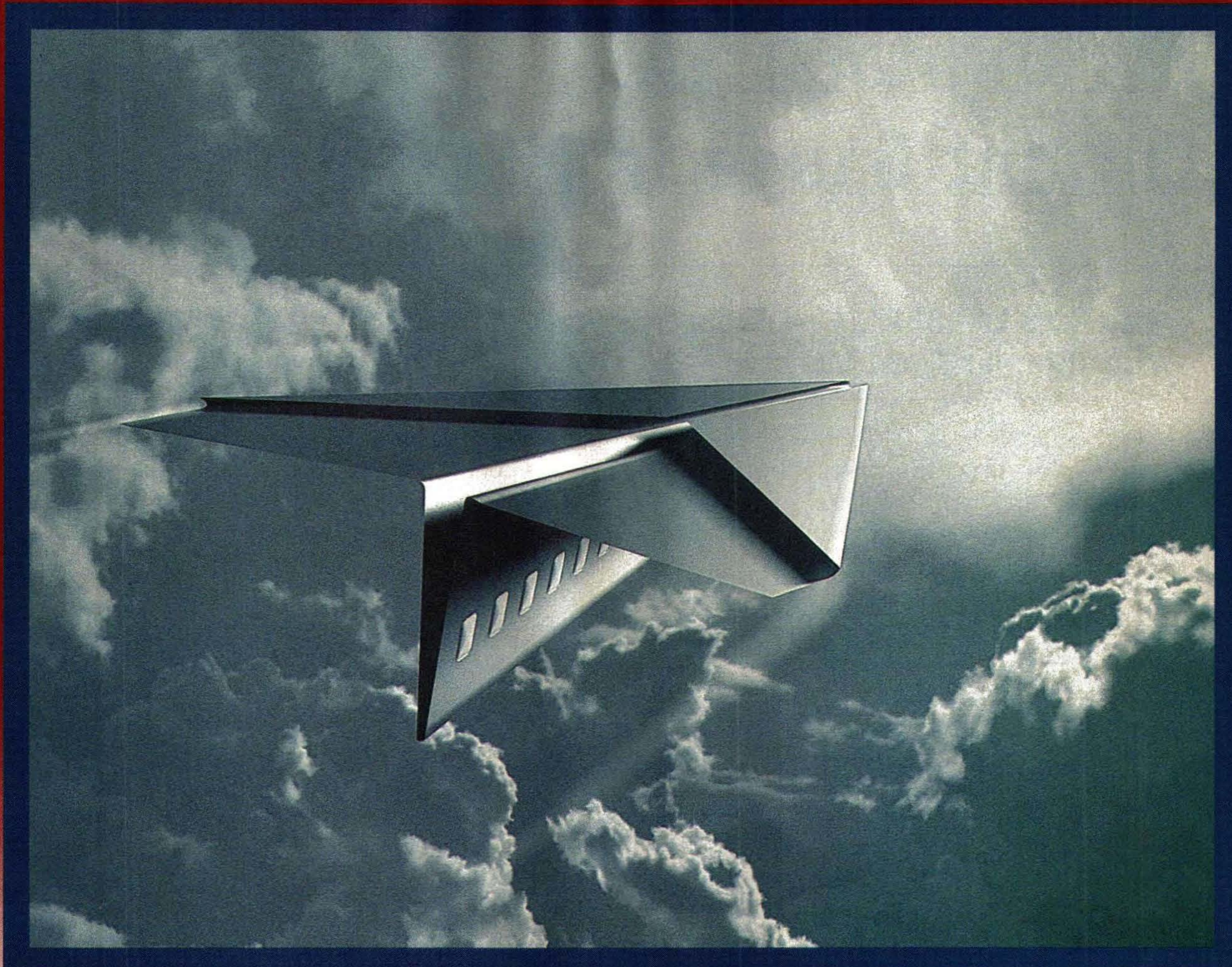
MINNEAPOLIS, MN—ADC recently announced the availability of the Digivance™ Indoor Coverage Solution (ICS). The Digivance ICS is an all-digital distributed-antenna system (DAS) that transports RF signals digitally within or between buildings. It is the initial product in ADC's Digivance family of digital RF transport solutions for managing and increasing wireless coverage and capacity.

The Digivance ICS is the only indoor coverage solution to offer a fully digital platform. To date, indoor coverage systems have used analog transport platforms only. By using patented technology, the Digivance ICS digitizes all signals in the fully allocated bandwidth, digitally transports them over multimode fiber, and reconstructs the signals at the far end.

"In our increasingly wireless world, users of mobile phones and other wireless devices expect a crisp and consistent signal regardless of their location," says Jeff Quiram, vice president and general manager for ADC's Broadband Connectivity Group, Wireless Division. "With the Digivance ICs, we've employed the latest in digital RF technology, providing the infrastructure needed to ensure clear wireless communications, including the higher data-rate services such as web access over a mobile device, within or between buildings."

Kudos

Reuben F. Richards, Jr., president and CEO of EMCORE Corp., has been inducted into New Jersey's High-Tech Hall of Fame. Richards has expanded EMCORE's business into four key growth-market areas: optical devices for broadband communications, electronic materials for wireless communications, solar cells for advanced satellite systems, and capital equipment for global communications and solid-state lighting applications.



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Model	Freq Range (GHz)	Vd (V)	Id (mA)	Gain (dB)	PldB (dBm)	IP_3 (dBm)	Thermal Resistance (°C/W)
NGA-186	0.1-6.0	4.1	50.0	12.5	14.6	32.9	120
NGA-286	0.1-6.0	4.0	50.0	15.5	15.2	32.0	120
NGA-386	0.1-5.0	4.0	35.0	20.8	14.5	25.8	144
NGA-486	0.1-6.0	5.0	80.0	14.8	18.3	39.5	118
NGA-586	0.1-6.0	5.0	80.0	19.9	18.9	39.6	121
NGA-686	0.1-6.0	5.9	80.0	11.8	19.5	37.5	121

Data at 1 GHz and is typical of device performance.

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NEWS

Semiconductor Processes

Si, GaAs, SiC, and SiGe expand designers' choices in power and small-signal wireless applications.

Semiconductors Vie For Space In Wireless Systems

Barry Manz
President
Manz Communications, 350 Main Rd., Montville, NJ 07045-9730; (973) 316-0999, e-mail: manzcom@erols.com.

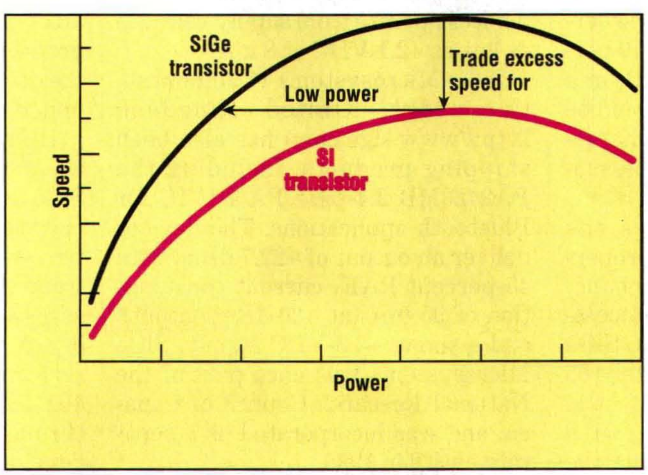
A technological whirlwind where advancements occur weekly is the wireless marketplace. In less than three years, digital signal processing (DSP) has gobbled up more of the functions in wireless phones, modulation techniques have grown from analog to complex digital, and development focus has shifted from voice to data. Designers are left to sort out the technologies that will drive their products into the market, and more often than not, the path to success is littered with the results of incorrect assumptions. Fortunately, development of semiconductor technology, although brisk, still leaves time for designers to choose the device best suited for a particular application.

Nevertheless, there are more competing processes and device types than ever before. When once there were silicon(Si)-based transistors, now there are Si, gallium-arsenide (GaAs), silicon-germanium (SiGe), silicon-carbide (SiC), and perhaps soon indium-phospate (InP) transistors, in device types ranging from bipolars to field-effect transistors (FETs), heterojunction bipolar transistors (HBTs), and high-electron mobile transistors (HEMTs). Within a few years, all of these devices may be vying for position in wireless handsets and infrastructure, which will make for lively competition. However, in practice, the current design environment supports a smaller number of devices and technologies, each one through its virtues and weaknesses having marked the limits of its design territory.

For example, Si laterally-diffused metal-oxide semiconductor (LDMOS) reigns supreme in base-station power amplifiers (PAs), with almost universal acceptance by base-station-amplifier manufacturers. The robust process has been thoroughly proven in a huge number of products for many years, and continues to advance in frequency and performance as it appears near its practical limit. Handset PAs continue to be well-served by GaAs HBTs, which have the third-order intercept performance required for current third-generation (3G) handsets. The GaAs HEMT has arguably some of the best noise performance found in current devices, and combined with its extended frequency range, beats back the challengers in noise-critical applications.

However, beyond these seemingly unassailable bastions lurk comparatively new technologies such as SiGe and SiC, which are already having impact in current designs. Of these newcomers, SiGe is the best developed and holds the greatest promise of displacing more traditional choices in some applications. In fact, although commercial SiGe devices have been available for only a few years, the process is being used in mainstream applications, thanks to the jump start provided by its compatibility with existing complementary-MOS (CMOS) processing facilities.

Development of SiGe has been spearheaded by IBM since 1982 and passed its first "technology qualification" at the company's Advanced



1. IBM's SiGe HBT delivers very-low noise figures, appealing for Rx applications.

Semiconductor Technology Center in Hopewell Junction, NY in 1996. It was originally envisioned by IBM as the successor to Si bipolar transistors for mainframe computers. However, dramatic changes in the mainframe market, increasing cost pressures, and development of the CMOS process combined to push SiGe into the weeds, if only for the moment.

However, the potential for SiGe was great enough for IBM to continue funding its development. It was long known that by modifying its bandgap, a bipolar semiconductor could achieve greater performance. To do this, it is necessary to grow an epitaxial layer of Ge-doped Si on the Si substrate, which in effect reduces the bandgap (the difference in energy state between a conducting and nonconducting electron in a device). This technique is the underlying driver that boosts three-to-four compounds, such as GaAs, to prominence in high-frequency applications. However, GaAs is a rare, expensive compound when compared to Si, which is ubiquitous and cheap by comparison.

Specifically, the lattice constraints of Si and Ge are somewhat different, and if Ge were grown on Si, a strain would result. This strain could be used to modify the bandgap and other properties of the material, effectively creating a new material with substantially higher electron mobility. Development of the process took years to achieve, but the resulting devices had performance more typical of compounds such as GaAs, with an f_{max} greater than 70 GHz, good noise performance, and power-added efficiency (PAE) up to 70 percent. Perhaps more important, in a practical sense, SiGe devices could be fabricated on existing, well-characterized, high-yield CMOS processing techniques (Fig. 1).

This effectively provides the resulting HBT device with properties akin to compound semiconductors such as GaAs, with the processing advantages of Si. That is, SiGe has a higher f_{max} (approximately 65 GHz) than Si alone, consumes little power, and is compatible with CMOS, so that a SiGe device may be fabricated on a Si chip alongside CMOS and bipolar. The device can accommodate more components on a single integrated circuit (IC), and deliver respectable efficiency and a low noise floor. It also provides a great degree of flexibility for designers of wireless circuits at 2.4 GHz and below, since the high-speed capability of the process can be traded off for better performance in other areas, such as lower power consumption, greater linearity, lower noise, or greater dynamic range (Fig. 2).

The processing advantages of SiGe are that new SiGe-based products frequently speed into the market. Two of the most recent are the SGA-0163 and SGA-0363 HBT monolithic-microwave IC (MMIC) amplifiers from Stanford Microdevices (Sunnyvale, CA, http://www.stanfordmicro.com) designed for use where low current consumption is a critical parameter, including wireless infrastructure and fixed wireless applications. They operate up to 5 GHz, and have small signal gain of 12 dB and 17 dB, respectively, at 2 GHz. Both devices operate from supply voltages as low as +2.1 VDC at 8 mA.

IBM HAS DEVELOPED THE FIRST GPS Rx CHIP SET WITH AN SiGe FRONT END. THE 12-CHANNEL Rx MEASURES 40 x 66 mm, BUT INCLUDES MEMORY, A GPS CRYSTAL, CONNECTORS, AND REAL-TIME CLOCK.

SiGe Microsystems (Cambourne, Cambridge, United Kingdom, http://www.sige.com) has also been shipping products, including the PA2423MB 2.4-GHz PA RF IC for Bluetooth applications. The devices deliver an output of +22.7 dBm, with 45-percent PAE, current consumption of 95 mA (at +20-dBm output), and a single +3.3-VDC supply. SiGe Microsystems was once part of the National Research Council of Canada, and was incorporated as a separate entity in 1996.

The MAX2645 LNA from Maxim Integrated Products (Sunnyvale, CA, http://www.maxim-ic.com) is designed for operation at 3.4-to-3.8-GHz wireless local loop (WLL), wireless broadband, and digital microwave radio applications. The device has typical gain of 14.4 dB, a +4-dBm input third-order intercept point (IP3), a 2.3-dB noise figure, and operates from +3.0 to +5.5 VDC at 9.2 mA. The device has a gain-step feature that reduces low-noise-amplifier (LNA) gain by 24 dB, while increasing input IP3 to +13 dBm, effectively improving receiver (Rx) front-end performance under high input-signal conditions, while reducing current consumption to 3 mA. IP3 is adjustable through an external bias resistor.

SiGe has also found a home in the prism chip set from Intersil (Mountaintop, PA, http://www.intersil.com). The company's ISL3685 2.4-GHz RF/IF converter and synthesizer includes a low-noise gain-selectable amplifier followed by a downconverter mixer in the receive chain, and an upconverter mixer and pre-amplifier for the transmit chain. The ISL3984 PA and detector include two stages, which together produce an output of +18 dBm. The detector is accurate over a dynamic range of 15 dB $\pm$1 dB. Both products operate from a single supply voltage of +2.7 to +3.6 VDC.

Atmel (San Jose, CA, http://www.temic-semi.de) recently announced that it has upgraded its SiGe foundry production capabilities in Heilbronn, Germany to a maximum frequency of 82 GHz. The company noted that after producing "millions" of SiGe products for its foundry customers, the decision was made to increase the upper-frequency limit.

IBM, the company that started it all, today offers products that rely on SiGe as well. The company has developed the first Global Positioning System (GPS) Rx chip set with a SiGe front end. The 12-channel Rx measures 40 $\times$ 66 mm, but includes memory, a GPS crystal, connectors, and real-time clock, as well as a PowerPC 401 embedded processor to boost throughput and enable application development. The direct-conversion Rx allows RF signals to be converted to digital signals without the need for

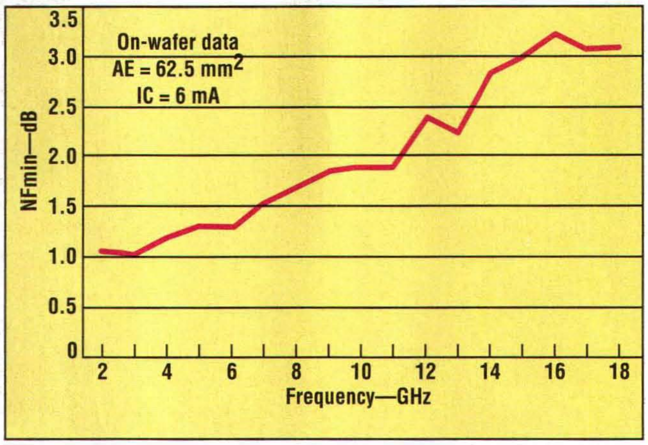
mixers, oscillators, and filters. The company also produces the IBM43RF0100, a SiGe negative-positive-negative (NPN) transistor that has a noise figure of 1.1 dB at 2 GHz, operating from +1 to +2.7 VDC, and has input IP3 capability of +10 dBm at 10 GHz. IBM also has relationships with organizations throughout the world to build devices using SiGe, and offers its SiGe fabrication facilities as a foundry service.

Si AND GaAs: STILL COOKING

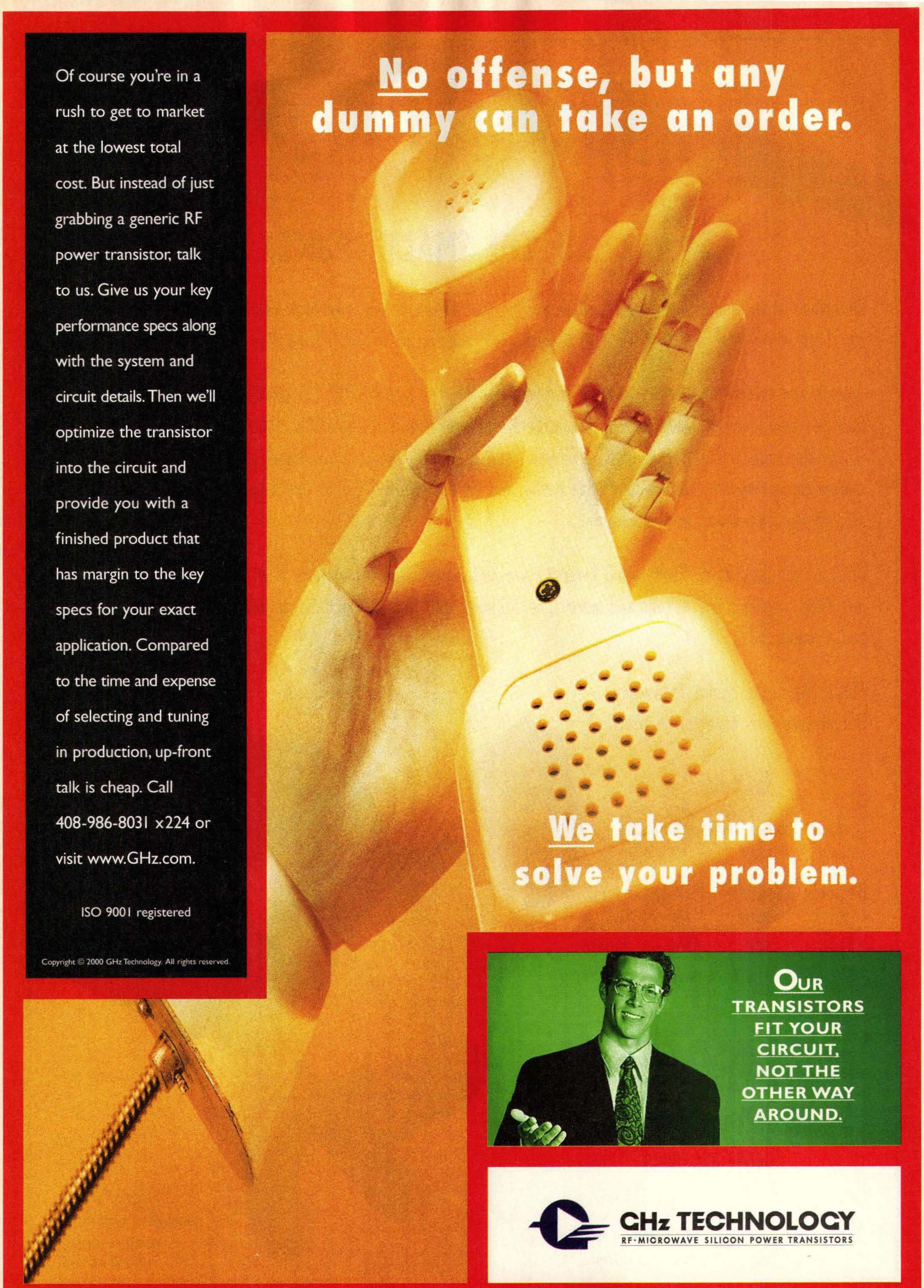
While a strong, amply-supported move is afoot to infuse current and future wireless applications with SiGe, the microwave industry has approximately 15 years of experience building systems around GaAs, and the number and types of GaAs devices dwarf those of SiGe. Proponents of GaAs argue that, when viewed as an overall system-building platform, GaAs has formidable advantages, including incorporation of passive components on the chip, a well-characterized process, and high yields. A true comparison between GaAs and alternatives, say its proponents, must include not only device cost, but the overall cost of designing it into a system. When this is done, GaAs competes far more favorably.

At Agilent Technologies (Santa Clara, CA, http://www.agilent.com), where GaAs continues to be king (at least for power applications) SiGe is being evaluated, according to David Wu, research-and-development (R&D) section manager for the wireless semiconductor division. "We feel that SiGe is very mature, and we're looking at applications for which it is best suited. We're implementing functions such as upconverters and downconverters in SiGe now, and the next generation of one of our chip sets (currently in Si) will be in SiGe."

The historical comparison



2. The high maximum oscillation frequency of SiGe can be traded off for reduced power consumption by reducing operating current.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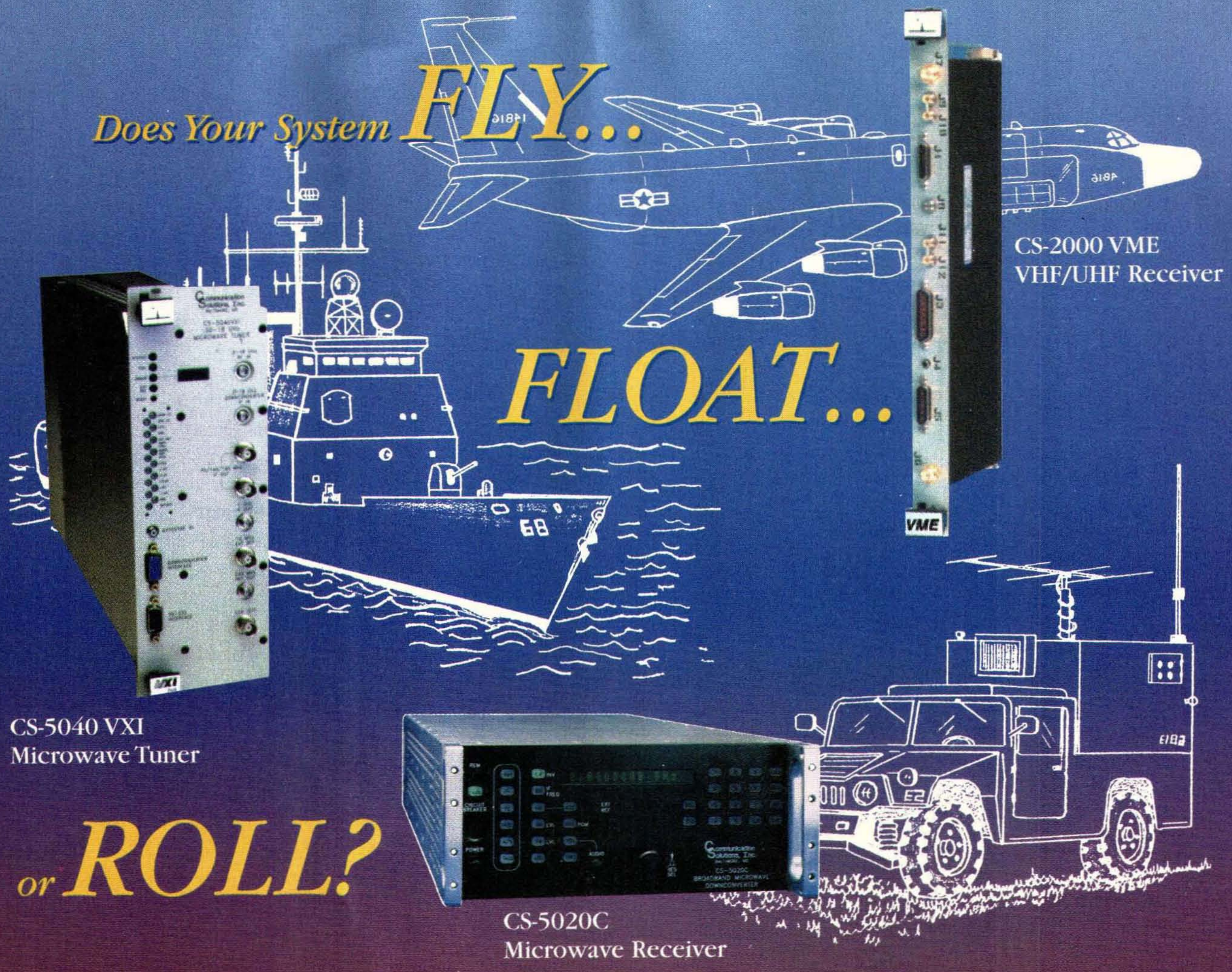
of FET versus bipolar for generation of RF power has created a hefty stack of white papers and application notes over the years. Motorola (Tempe, AZ, http://www.motorola.com), bastion of the Si bipolar, has contributed much of this material, and today manufactures both types of devices, covering virtually any likely application, from low-power wireless handsets to high-power radar transmit/receive (T/R) modules, medical systems, and high-power high-frequency (HF), very-HF (VHF), and ultra-HF (UHF) transmitters (Txs). The LDMOS FET is likely to be broadly employed in wireless base stations as it is today, thanks to its robust nature and favorable electrical characteristics.

A relative newcomer to the power market comes from Agilent, which is investing heavily in its GaAs enhancement-mode pseudomorphic HEMT (E-PHEMT) process. This technology, designed for the low-power amplifiers of wireless handsets, requires no negative voltage supply, unlike GaAs MESFETs and HEMTs. The company believes that PA modules using this technology can deliver a 15-percent increase in battery life, and reduce manufacturing cost and board space by eliminating the negative supply. Agilent recently announced plans to build a 15,000 ft.^{2} clean-room facility in Ft. Collins, CO, ranging from Class 100 to Class 1, and dedicated to producing up to 48,000 six-in. (15.24-cm) wafers per year in support of its Global System for Mobile Communications (GSM) and code-division-multiple-access (CDMA) PA business.

PCS WIRELESS SYSTEMS CURRENTLY EMPLOY MILLIMETER-WAVE RADIOS AT FREQUENCIES UP TO 38 GHz TO IMPLEMENT THE BACKHAUL LINKS FROM BASE STATIONS.

The millimeter-wave region above 18 GHz has always been viewed as the next big thing for commercial applications, although until very recently, just what applications these would be was subject to extreme speculation. However, the extraordinary acceptance of the Internet and personal wireless communications may soon allow even the most conservative pundits to safely predict real markets in this little-used region of the spectrum.

Personal-communication-services (PCS) wireless systems currently employ millimeter-wave radios at frequencies up to 38 GHz in order to implement the backhaul links from base stations, a market that has generated sizable revenues for manufacturers. In addition, the first adaptive cruise-control systems (once known as collision-avoidance radar) are find-



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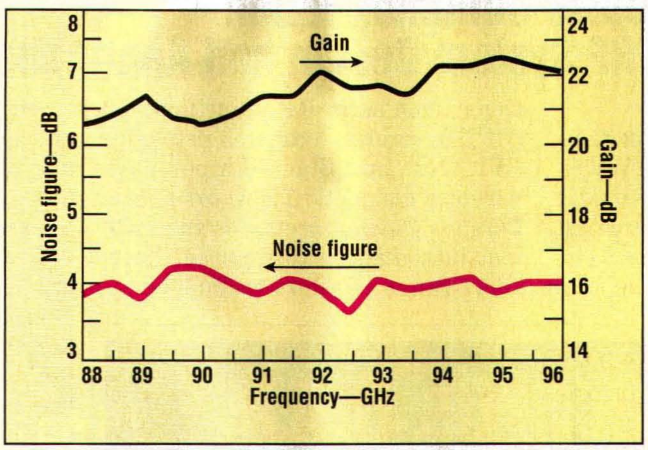
ing their way into luxury automobiles. These systems, which operate at 77 GHz, have been tested in the field for years, but only recently have been integrated with the signal processing and computational horsepower needed to make split-second decisions.

In addition, the local-multi-point-distribution-system (LMDS) concept shows signs of being able to compete with wired solutions such as cable modems and asymmetric digital subscriber line (ADSL) for delivering high-speed Internet access and perhaps other applications. Operating between 27 and 30 GHz, LMDS will require inexpensive low-power microwave transceivers in large numbers. Wireless Ethernet, proposed for operation between 55 and 60 GHz, may ultimately become reality as well.



3. This W-band three-stage InP MMIC LNA developed at HRL Laboratories shows good noise figure and gain performance.

One semiconductor compound that has long held promise at millimeter-wave frequencies up to 100 GHz is InP. At these frequencies, the InP HEMT is the only potentially viable option that can also deliver the large-scale integration required in these applications. InP is currently employed in military systems in the form of HEMTs. Research is currently underway at companies such as HRL Laboratories, Lucent Technologies, TRW, and Nortel, as well as at universities and companies such as Hitachi, NEC, and NTT, to transform InP technology from its limited use in military systems to a cost-effective, commercially-viable solution for high-speed data and signal-processing applications (Fig. 3).

Progress in making this transition is currently limited to the high cost of a typical 7.5-cm wafer, and high-defect densities occurred during material growth and processing. However, the almost certain need for extremely high-speed processing capabilities in many applications makes it likely that withinthe decade, InP development will advance to the point at whichcommercial products can be manufactured. ••

Short-Range Wireless Communication Fundamentals of RF System Design and Applications

Alan Bensky

Wireless communications is perhaps best known by its "long-range" applications, such as cellular and satellite communications. Perhaps the greatest growth ahead for wireless markets, however, lies in "short-range" applications, such as in RF identification (RFID), wireless local-area networks (WLANs), and Bluetooth personal wireless connectivity. *Short-range Wireless Communication* is an excellent introduction to some of these short-range wireless technologies, especially for engineering managers and marketing professionals who may not be well-versed in the terminology and technology of wireless communications.

The opening chapter offers a brief historical perspective on wireless communications, along with an introduction to generic wireless communications systems. Chapter 2 introduces the concept of radio propagation. Chapter 3 highlights antennas and transmission lines.

Chapter 4 reviews communication protocols and modulation. Chapter 5 covers transmitters (Txs), using a simple block diagram to discuss the roles of different components, such as the modulation source and the amplifier, while Chapter 6 details receivers (Rxs), with a short section on software-defined radios.

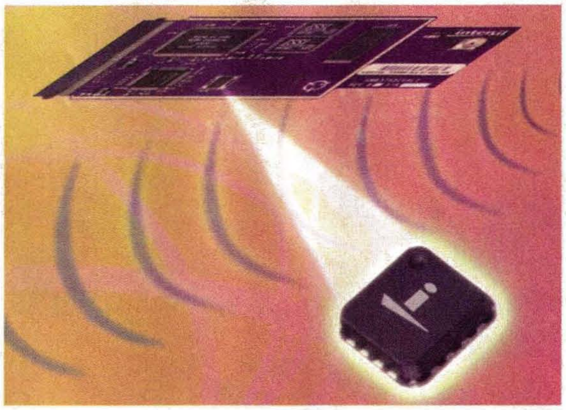
Chapter 7 unveils radio-system design, with a step-by-step discussion on key system parameters, such as operating range, sensitivity, noise figure, and bandwidth, as well as how to calculate performance when provided with different values of these parameters. Chapter 8 covers system implementation, using module and integrated-circuit (IC) products from a variety of manufacturers as examples. Chapter 9 reviews regulations and wireless standards.

Chapter 10 is a brief 27-page introduction to information theory, with coverage of probability-density functions (PDFs). Chapter 11 details new developments in short-range radios, including brief treatments of ultra-wideband radios and Bluetooth technology.

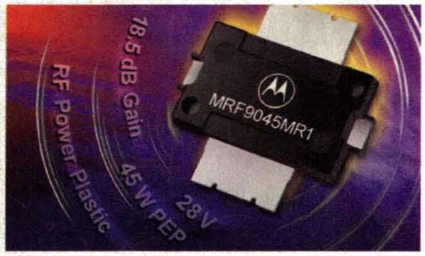
Short-range Wireless Communication is accompanied by a compact-disc read-only memory (CD-ROM) that contains MathCAD worksheets on 12 topics covered in the text (such as patch-antenna design). (2000, 300 pp., paperback, ISBN: 1-878707-53-1, $49.95.) **LLH Technology Publishing, 3578 Old Rail Rd., Eagle Rock, VA 24085; (800) 247-6553, (540) 567-2000, FAX: (540) 567-2539, e-mail: carol@LLH-Publishing.com, Internet: http://www.LLH-Publishing.com.**

Amp/detector targets WLANs

The model ISL3984 integrated circuit (IC) contains a power amplifier (PA) and a detector and operates in the 2.5-GHz band for wireless-local-area-network (WLAN) applications. The chip is designed to be used in conjunction with the company's PRISM 2.5 WLAN chip set. It is said to maximize battery life while delivering the power necessary for high-performance, 11 Mb/s wireless networking systems. The amplifier section offers a typical power gain of 30 dB and a typical RF output power of +18 dBm. The detector section has a logarithmic power-detect function that provides a DC output voltage proportional to the logarithm of the output power. The slope of the detector output voltage is 100 mV/dB over a 15-dB dynamic range. The chip has a complementary-metal-oxide-semiconductor (CMOS)-compatible power up/down function and is said to draw 25 percent less current than its predecessor. It is housed in a 16-lead, multilead flat pack (MLFP). **Intersil Corp., 125 Crestwood Rd., Mountaintop, PA 18707; (888) 468-3774, Internet: http://www.intersil.com.**



For more information, visit www.mwrf.com



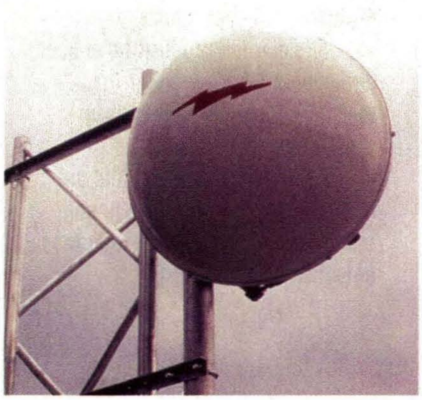
Power MOSFET serves base stations

The model MRF9045MR1 metal-oxide-semiconductor field-effect transistor (MOSFET) is designed for broadband commercial and industrial applications at frequencies to 1 GHz. It is ideal for large-signal, common-source amplifier applications in +28-VDC base-station equipment. The laterally-diffused-metal-oxide-semiconductor (LDMOS) N-channel device can deliver up to 45 W of peak envelope power (PEP). At 945 MHz, its typical power gain is 18.5 dB and its two-tone efficiency is typically 41 percent. Typical intermodulation distortion (IMD) is 31 dBc. The transistor is available in a TO-270 two-lead plastic package designed to be compatible with automated printed-circuit-board (PCB) assembly lines. **Motorola Semiconductor Products Sector, P.O. Box 52073, Phoenix, AZ 85072-2073; (602) 413-4991, FAX: (602) 413-7986, Internet: http://www.motorola.com/sps.**

For more information, visit www.mwrf.com

Antennas serve point-to-point applications

The ValuLine series of terrestrial microwave antennas now includes 2.2- and 1.8-m diameter models available for three frequency bands: 5.925 to 6.425 GHz, 6.425 to 7.125 GHz, and 5.925 to 7.125 GHz. The 5.925- to- 6.425-GHz antennas are available in two unshielded, single-polarized versions and in a shielded, single-polarized version. The 6.425-to-7.125-GHz antennas are available in three unshielded,

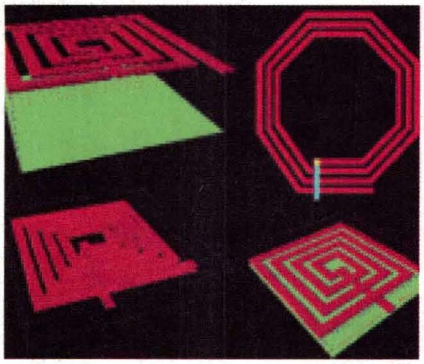


single-polarized versions, a shielded, single-polarized version, and a shielded, dual-polarized version. The 5.925-to-7.125-GHz antennas are available in unshielded and shielded, single-polarized versions. **Andrew Corp., 10500 W. 153rd St., Orland Park, IL 60462; (708) 349-3300, FAX: (708) 349-5222, Internet: http://www.andrew.com.**

For more information, visit www.mwrf.com

Software helps design printed spiral inductors

The SPIRAL inductor analysis and design software package helps engineers design, characterize, and optimize spiral inductors on silicon gallium arsenide (Si GaAs),



ceramic, and printed-circuit boards for wireless communications. The package uses field simulations to output fully coupled, optimally sectioned remote line concentrating module (RLCM), S-parameter, and Z-parameter models. It generates single-metal, double-metal, and differential spirals. The program's calculations includes skin and substrate effects, and builds a GDSII layout of the finished spiral design. **OEA International, Inc., 3235 Kifer Rd., Suite 300, Santa Clara, CA 95051; (408) 738-5972, FAX: (408) 738-2017, Internet: http://www.oea.com.**

For more information, visit www.mwrf.com

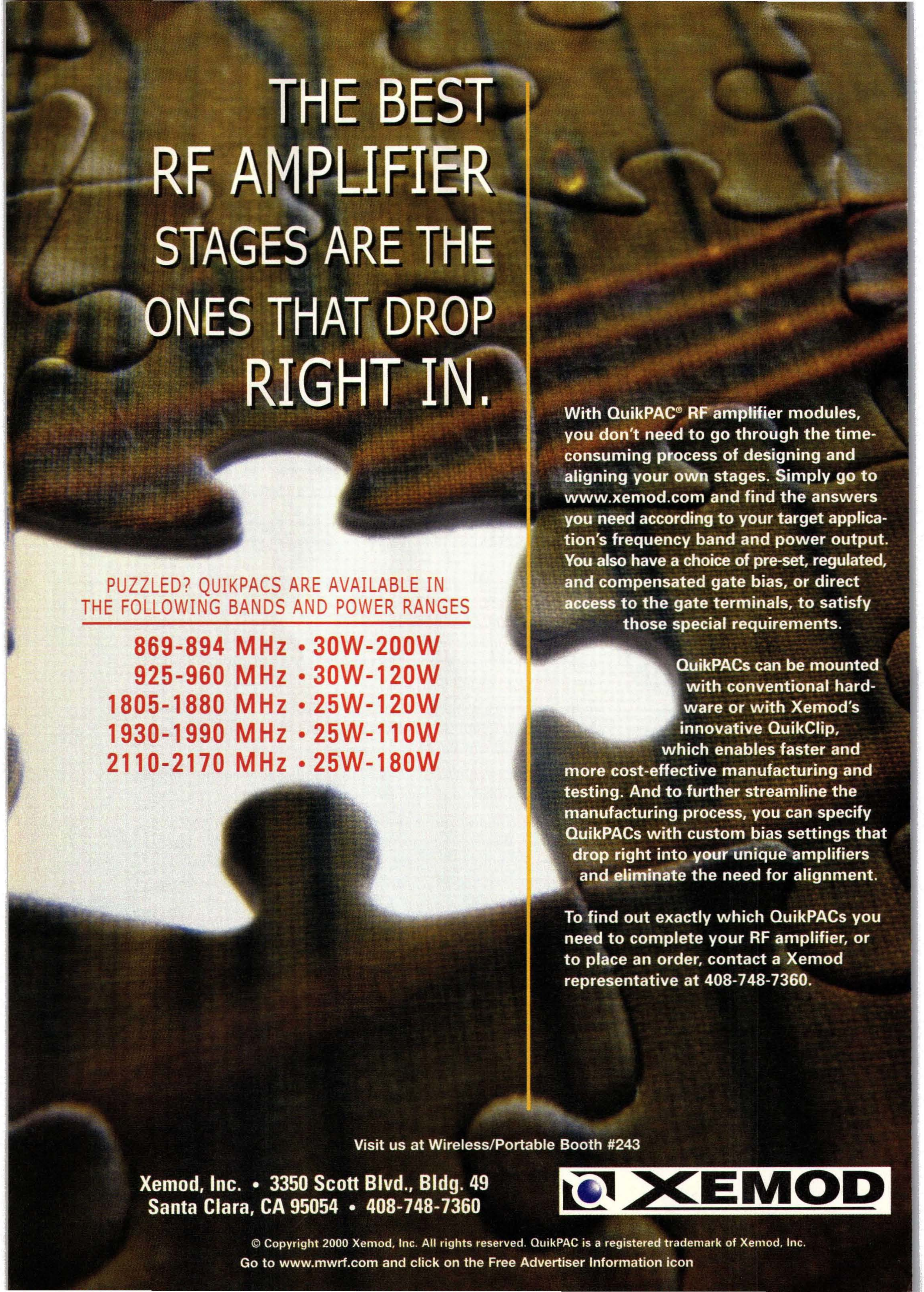
Cree Strikes Again In RF Power Amps

In another foray into the RF linear power-amplifier (PA) infrastructure market, Cree, Inc. (Durham, NC) entered into an agreement with Spectrian Corp. (Sunnyvale, CA) that gave Cree all of the assets of Spectrian's UltraRF power-semiconductor division. Just three months ago, Cree and Xemod (Santa Clara, CA) joined forces to develop Cree's silicon-carbide (SiC) power-semiconductor technology for wireless base-station applications (see "Cree, Xemod to Jointly Develop SiC," *Microwaves & RF*, November 2000, p. 40).

Under the latest agreement, Cree purchased the assets of Ultra RF and assumed certain liabilities in exchange for approximately 1,816,600 shares of Cree common stock with additional Cree shares worth $30 million. A second part of the transaction is that the companies entered into a two-year agreement for Cree to supply RF power semiconductors to Spectrian. A third provision of the deal is a joint agreement for the two companies to develop advanced technologies related to laterally diffused metal-oxide semiconductors (LDMOS), high-linearity and gain-driver modules, high-efficiency LDMOS power modules, and SiC metal-semiconductor field-effect transistors (MESFETs).

UltraRF was wholly owned by Spectrian and boasts approximately 125 employees. Its specialty is bipolar and LDMOS power transistors and modules for the wireless infrastructure. Revenues are approximately $7 to $8 million, virtually all of it to Spectrian. As part of Cree, Ultra RF will have a greater opportunity to grow its external sales, leaving Spectrian to focus on its primary market, building single- and multicarrier linear, high-power RF amplifiers for the worldwide wireless base stations.

For Cree, this second venture into the PA and semiconductor markets represents an effort to become a major supplier to wireless infrastructure suppliers such as Spectrian and others. At present, LDMOS power devices are favored for most wireless telephone applications up to 2 GHz. However, as wireless technology moves into its third generation (3G) and beyond, semiconductor and infrastructure manufacturers must plan for equipment that will operate at frequencies in the range of 2 to 5 GHz. Operation in this range may be beyond the limits for LDMOS, so alternatives must be found. While the full potential of SiC as a power-semiconductor technology is yet to be realized, it offers many technical advantages over LDMOS and other power technologies, including the ability to operate in the 5-GHz region. Since Cree is a leader in SiC, but not high-frequency PAs, it is forming alliances to establish a foothold in the future infrastructure market. ••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RF AMPLIFIER
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PUZZLED? QUIKPACS ARE AVAILABLE IN
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925-960 MHz • 30W-120W
1805-1880 MHz • 25W-120W
1930-1990 MHz • 25W-110W
2110-2170 MHz • 25W-180W
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QuikPACs can be mounted with conventional hardware or with Xemod's innovative QuikClip, which enables faster and more cost-effective manufacturing and testing. And to further streamline the manufacturing process, you can specify QuikPACs with custom bias settings that drop right into your unique amplifiers and eliminate the need for alignment.
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Contracts

Wireless Technologies Corp.—Has accepted an order for the supply of several thousand diplexers from Cellular Specialties, Inc. (Manchester, NH). These diplexers will be employed in CSI's specialized-mobile-radio (SMR) high-performance repeaters and cellular extenders.

Teradyne, Inc.—Announced that Dallas Semiconductor (Dallas, TX), a complementary-metal-oxide-semiconductor (CMOS) chip manufacturer, has purchased multiple Catalyst and INTEGRA J750 test systems for mixed-signal telecom and microcontroller device testing. The total order exceeded $5 million.

Telefonica Moviles—Has awarded Motorola the contract to supply general-packet-radio-services (GPRS) high-speed mobile data services on the Movistar Global System for Mobile Communications (GSM) network. Commercial launch of the GPRS core network, enabled by Motorola and strategic-alliance partner Cisco Systems, Inc., is expected by the end of this year.

Gabriel Electronics, Inc.—Was recently awarded a contract to supply 1000 26-GHz subscriber antennas to Israeli system provider Floware Wireless Systems Ltd. (Or Yehuda, Israel). This award is the first stage of a much larger contract. Gabriel has developed and provided to Floware an ETSI-certified subscriber antenna designed for direct-connect integration into the Floware WALKair™ radio.

Alpha Industries—Announced that it is delivering additional volume shipments of gallium arsenide (GaAs) and RF components for Metricom, Inc.'s high-speed Ricochet™ mobile data network. Ricochet provides 128-kb/s wireless Internet access to users in nine metropolitan areas across the US, and is under construction in approximately one dozen additional markets.

M/A-COM SIGINT Products—Has been awarded a contract by the Naval Surface Warfare Center (NSWC) [Virginia Beach, VA] to supply a large quantity of collection microwave receivers (Rxs) over a five-year period.

Rockwell Collins—Has been awarded an engineering-services contract by the US Air Force to provide continued technical support of the avionics systems installed on its fleet of C/KC-135 aircraft. As part of the contract, Rockwell Collins will provide engineering support for the entire C/KC-135 avionics suite, which includes all equipment installed under the Pacer CRAG (Compass, Radar, and Global Positioning System) and GATM (Global Air Traffic Management) programs, as well as the legacy avionics systems.

Fresh Starts

UltraRF—Has selected Sunrise Technology as its distributor in China. Sunrise Technology has more than 20 sales offices located in the People's Republic of China, including offices in Shanghai, Beijing, and Shenzen.

Applied Wave Research, Inc. (AWR)—Signed a comprehensive agreement with Global Communication Semiconductors, Inc. (GCS) to provide electronic-design-automation (EDA) solutions to be used on a worldwide basis for the design of microwave integrated circuits (ICs). GSC is a provider of heterojunction bipolar transistors (HBTs), pseudomorphic-high-electron-mobilty transistors (PHEMTs), epitaxial wafers, surface-acoustic-wave (SAW), and optoelectronic devices.

Ericsson Microelectronics—Has begun production of its DC-to-DC converters at a newly acquired site at Kalmar, Sweden. New production lines have been set up at the former Volvo factory. With further planned expansion, output of the company's DC-to-DC converters will be tripled by the end of 2001.

Flarion Technologies—Has relaunched its corporate website. Offering an improved user interface with easy-to-follow links, Flarion's website provides information about the company's Flash-OFDM technology, which promises to provide affordable wireless Internet access and significant technological advantages over current and third-generation (3G) wireless solutions. Visitors to Flarion's website can read the latest company news, gain knowledge of the latest trends in the wireless industry, as well as learn about employment opportunities with Flarion. The URL of Flarion's website is http://www.flarion.com.

AEMC Instruments—Is moving to a modern facility in Foxborough, MA following two decades in downtown Boston. The new office will accommodate AEMC's recent growth and provide room for their continued expansion. Additionally, the larger facility will provide better access to technical resources, as well as convenient access to major highways and airports. AEMC's US-based research-and-development (R&D) team will join the sales and marketing team, thus making this new location key to their continued product success. AEMC's manufacturing and distribution operations will remain in their Dover, NH facility.

Interconnect Devices, Inc.—Announced the redesign and expansion of their website, which can be accessed at http://www.idinet.com. Launched in 1994, idinet.com was one of the first websites in the automatic-test-equipment (ATE) industry.

Lucent Technologies—Announced that its Microelectronics spinoff has selected Agere Systems as its name. The company, which is the former Microelectronics Group of Lucent Technologies, is the world leader in semiconductors for communications applications and is comprised of two major divisions—integrated circuits (ICs) and optoelectronic components.

Labtech Ltd.—Announced that it is trebling its plating capacity. Labtech recently installed a new electrolytic nickel (Ni) and hard and soft gold (Au) line, a new fully automatic electroless NiAu line, and a new hot-air solder-levelling process. They have also renewed their resist-strip chemical-etch line to a higher specification. A new electrolytic plating line from PAL and a new oxide coating line have been in use since the end of September 2000. The investment, totaling 750,000 pounds (approximately $1,087,500 US), completes Labtech's major capital expenditure program in the chemical-processing area and is the end of a two-year program to provide a higher technical tolerance capability with automated SPC production.

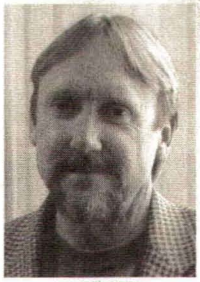
A ________

B ________

C ________

room33—Huw Hampson-Jones to CEO; formerly executive vice president at Sonera's iD2 Technologies.

TriPoint Global's VertexRSI—Anthony D. "Tony" Radford to vice president of system sales for the Duluth, GA business unit; formerly director of system sales for the satellite-systems business unit.



RADFORD

HAN

ITT Industries, Cannon—Lee Han to private branding manager; formerly staff project engineer for Bourns, Inc.

Enthone—Stephen LaCroce to vice president of marketing for Enthone Performance Coatings; formerly vice president of marketing for Enthone PWB Chemistries. Also, Raymundo Gonzalez to general manager for South America; formerly commercial director for Enthone de Mexico. In addition, C.W. Law to vice president for Asia; formerly managing director for Greater China.

AirDesk, Inc.—Mark Panetta to CFO; formerly CFO at V-Comm.

CTS Corp.—David T. Ciembronowicz to director of worldwide sales for CTS Wireless Components, Inc.; formerly vice president of sales and marketing for Mhotronics, Inc. Also, Alan B. Bennett to director of sales and marketing for the CTS Reeves Frequency Products business unit; formerly director of sales and marketing with Amphenol Fiber Optic Products.

Scott Specialty Gases—Lois J. Hayes to corporate controller; formerly director of accounting.

Andrew Corp.—Dr. Rolf Bergmann to vice president for Europe, Africa, Middle East, South Asia, and C.I.S.; formerly managing director of European sales. Also, Jim McIlvain to vice president for Asia Pacific and global OEM sales; formerly director of Asia Pacific and global OEM sales. In addition, Donn Peterson to vice president of wireless infrastructure sales for the US and Canada; formerly director of wireless infrastructure.

Signal Technology Corp.—Dr. James G. Oakes to vice president of cellular/PCS/Wireless Data Products; formerly acting deputy general manager at Raytheon's RF components division.

Balzers Thin Films, Inc.—Richard T. Seery to senior sales engineer; formerly international sales manager at Villa Precision.

Intertek Testing Services (ITS)—Salvatore Napoli to operations manager for the ETL SEMKO, Americas division's Boxborough, MA facility; formerly operations manager for General Electric's Industrial Systems Division in Plainville, CT.

Seiko Instruments USA, Inc.—Steve Baldo to general manager of the Optical Fiber Components Group, Electronic Components Division; formerly national sales manager.

JMS Worldwide, Inc.—Sridhar Kowdley to CTO; formerly director of RF engineering.

Cooper Electronic Technologies—Steven Hogge to vice president and general manager; formerly vice president of sales.

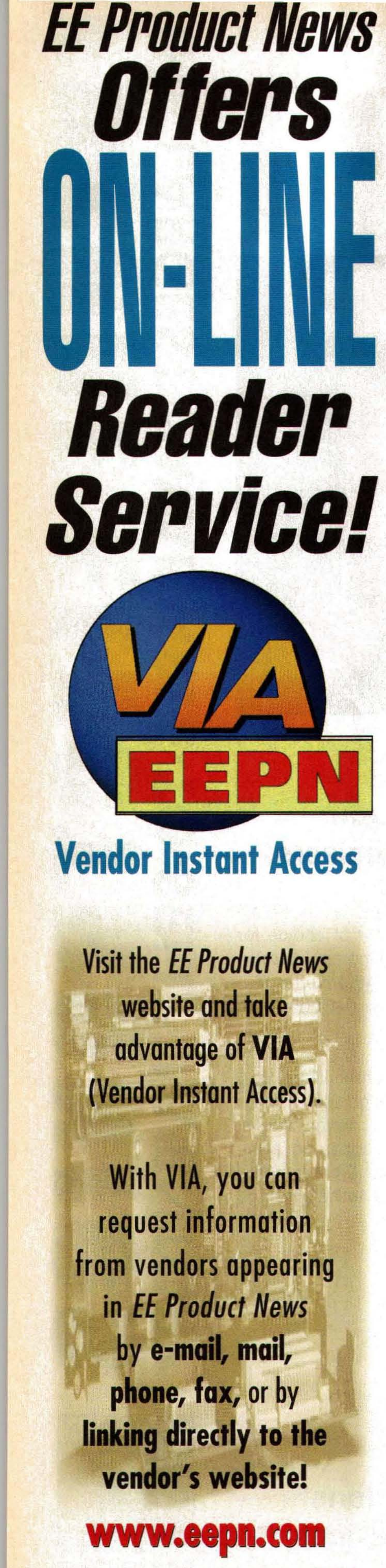
HOGGE

POUCHET

Racal Instruments—Jack Pouchet to the position of marketing director; formerly director of sales and marketing at Switching Systems International.

Narda Microwave-East—Thomas O'Rourke to regional sales manager; formerly international applications engineer.

IPC—Clint Gendusa to media-relations manager; formerly worked in PR and promotions for Group III Promotions.



Short Courses

Short Range Wireless and Bluetooth
January 31-February 2 (Mountain View, CA)
Besser Associates
201 San Antonio Circle, Bldg. E
Mountain View, CA 94040
(650) 949-3300, FAX: (650) 949-4400
e-mail: info@bessercourse.com

Telecommunications Fundamentals I and II
February 20-23 (Anaheim, CA, Houston, TX, Minneapolis, MN)
Global Knowledge
9000 Regency Pkwy., Suite 500
Cary, NC 27511
(800) COURSES, FAX: (919) 461-8646
Internet: http://am.globalknowledge.com

Using and Installing Fiber Optic Systems for Communications
February 26-28 (Madison WI)
Dept. of Engineering Professional Development
University of Wisconsin-Madison
Madison, WI 53706
(800) 462-0876, FAX: (608) 263-3160
e-mail: custserv@epd.engr.wisc.edu
Internet: http://epd.engr.wisc.edu

Practical Design of Integrated and Discrete Wireless Circuits
February 26-28 (Mountain View, CA)
Besser Associates
201 San Antonio Circle, Bldg. E
Mountain View, CA 94040
(650) 949-3300, FAX: (650) 949-4400
e-mail: info@bessercourse.com

Wireless R&D Symposium
March 6 (Boca Raton, FL)
March 8 (Piscataway, NJ)
March 13 (Boston, MA)
Agilent Technologies
(800) 752-1545
Internet: http://www.agilent.com/find/wirelessRDsymp

Hands-On Fiber-Optic Systems
March 6-9 (Washington, DC)
Learning Tree International
1805 Library St.
Reston, VA 20190-5630
(800) 843-8733, FAX: (800) 709-6405
Internet: http://www.learningtree.com

Wireless Digital Communications
March 12-16 (Tempe, AZ)
Arizona State University
Center for Professional Development
Tempe, AZ 85287-7505
(480) 965-1740, FAX: (480) 965-8653
e-mail: asu.cpd@asu.edu
Internet: http://www.eas.asu.edu/cpd

Meetings

DesignCon 2001
January 29-February 1 (Santa Clara Convention Center, Santa Clara, CA)
DesignCon 2001 (International Engineering Consortium)
549 West Randolph St., Suite 600
Chicago, IL 60661-2208
Internet: http://www.iec.org

International Solid State Circuits Conference (ISSCC)
February 5-7, 2001 (San Francisco Marriott Hotel, San Francisco, CA)
The Solid State Circuits Society of the IEEE
Piscataway, NJ 08855-1331
Internet: http://www.isscc.org

Commercialization of Military and Space Electronics Conference & Exhibition
February 12-15, 2001 (Los Angeles Airport Marriott Hotel, Los Angeles, CA)
Components Technology Institute, Inc.
904 Bob Wallace Way
Suite 117
Huntsville, AL 35801
(256) 536-1304, FAX: (256) 539-8477
e-mail: dale@cti-us.com
Internet: http://www.cti-us.com

Mobile Commerce Conference
February 12-13 (Fairmont Hotel, San Jose, CA)
Keye3Media Events
Needham, MA 02494
(800) 482-8144
Internet: http://www.key3media.com/mc2

Wireless/Portable Symposium & Exhibition
February 12-16 (San Jose Convention Center, San Jose, CA)
Penton Media, Inc.
Wireless/Portable 2001
611 Route 46 West
Hasbrouck Heights, NJ 07604
FAX: (201) 393-6297
Internet: http://www.WirelessPortable.com

Wireless Internet—Impact on Next Generation Handheld Product Packaging
February 18-20 (Boscolo Park Hotel, Nice, France)
International Wireless Packaging Consortium (IWPC)
Don Brown, Director
(215) 491-2113, FAX: (215) 491-2115
e-mail: donbrown@iwpc.org
Internet: http://www.iwpc.org

Intel Developer Forum Conference, Spring 2001
February 26-March 1 (San Jose Convention Center, San Jose, CA)
Intel Developer Forum
110 Shawmut Rd.
Canton, MA 02021-1411
(888) 397-6205, FAX: (781) 828-9992
Internet: http://developer.intel.com/idf

CTIA Wireless 2001
March 20-22 (Sands Expo & Venetian Hotel, Las Vegas, NV)
CTIA Wireless 2001, c/o CMR
33 New Montgomery, 14th Fl.
San Francisco, CA 94105
(800) 421-6338, FAX: (415) 979-2250
Internet: http://www.ctiashow.com

CeBit
March 22-28 (Hannover Fairgrounds, Hannover, Germany)
Hannover Fairs USA, Inc.
103 Carnegie Center
Princeton, NJ 08540
(609) 987-1202, FAX: (609) 987-0092
e-mail: info@hfusa.com
Internet: http://www.hfusa.com

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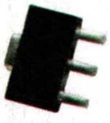
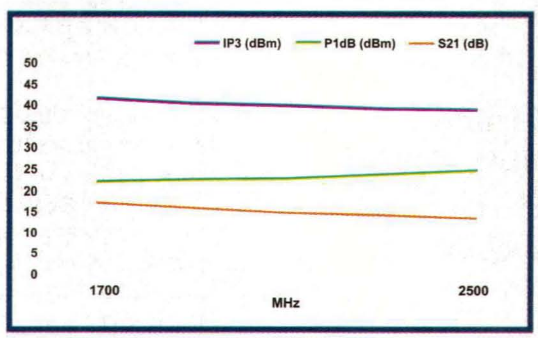
Introducing the SXT-289. A new wideband GaAsHBT power amplifier.

Stanford Microdevices introduces the SXT-289 — the perfect driver amplifier for today's and tomorrow's advanced communication infrastructure equipment.



With power to spare, giving you freedom from design constraints, this amplifier was designed and manufactured using state-of-the-art Gallium Arsenide heterojunction bipolar transistor (GaAs HBT) process technology. These devices are ideal for use as a driver stage for power amplifiers installed in cellular PCS infrastructure equipment and conform to TDMA, CDMA and PCS 1900 modulation standards. The SXT-289 operates from a single 5VDC supply and offers exceptional linearity performance in a small form-factor plastic SOT-89 with backside ground.

For more information on the SXT-289 or any of our other products, visit our website today and experience RF Innovation from Stanford Microdevices.



Typical device performance. Bias =5V @ 110mA typ.

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SiGe/Si challenges AlGaAs/GaAs in handset HBT amp

For the past several years, the overwhelming majority of heterojunction-bipolar-transistor (HBT) power amplifiers (PAs) used in dual-mode, code-division-multiple-access/Advanced Mobile Phone Service (CDMA/AMPS) handset transmitters (Txs) have been manufactured of aluminum gallium arsenide (AlGaAs) and GaAs because they have excellent linearity and power-added efficiency (PAE). However, silicon-germanium/silicon (SiGe/Si) HBT amplifiers are more attractive because they achieve comparable performance with lower emitter-base turn-on voltage, higher thermal conductivity, and lower production cost. IEEE Senior Members Pei-Der Tseng, Liyang Zhang, Guang Bo Gao, and IEEE Fellow M. Frank Chang have successfully designed and built an HBT amplifier integrated circuit (IC) for dual-mode handsets based on a standard SiGe/Si HBT foundry technology. The +3-VDC chip measures 2 × 1 mm and meets all linearity and output-power requirements for handset operation. See "A 3-V Monolithic SiGe HBT Power Amplifier for Dual-Mode (CDMA/AMPS) Cellular Handset Applications," *IEEE Journal of Solid State Circuits*, September 2000, Vol. 35, No. 9, p. 1338.

TCBs might save manufacturers millions

When a telecommunications manufacturer develops a new piece of equipment, the device must be tested and approved by the Federal Communications Commission (FCC) before it can be marketed. But this process can take anywhere from 40 to more than 100 days, and some manufacturers have estimated that they can lose up to $1 million a day waiting for FCC approval. According to Donald L. Sweeney of D.L.S. Electronic Systems, Inc., a group of Telecommunications Certifications Bodies (TCBs) will soon function as extensions of the FCC, reducing this part of the FCC's workload, speeding the test-and-approval process, and saving manufacturers money. To become a TCB, an organization must have the technical expertise to understand and review technical and administrative applications submitted by manufacturers. To this end, the FCC sponsored a five-day workshop in December of 1999 to train potential TCBs. The FCC has also mandated a TCB Council to ensure that TCBs perform their duties in a uniform manner. See "Telecommunication certification bodies—Questions and Answers," *Interference Technology Engineers' Master ITEM 2000*, p. 20.

TFR measurements simplify GSM equipment tests

Evaluating Global System for Mobile Communications (GSM) equipment involves various types of test instruments that are often very sophisticated and expensive, and often require specific features and elaborate operations to successfully analyze GSM signals in a single domain (either time or frequency). However, three Italian researchers propose a digital-signal-processing (DSP) approach to testing GSM signals that takes advantage of the intrinsic time-frequency nature of GSM signals, thus making the tests more agile, entirely automatic, and time and cost-effective. Leopold Angrisani of the University of Napoli Federico II, Pasquale Daponte of the University of Sannio, and Massimo D'Appuzo of the University of Napoli demonstrated this technique, which performs measurements on GSM signals in the time and frequency domains simultaneously by downconverting, digitizing, and converting the signals into time-frequency representations (TFRs). See "A Measurement Method Based on Time-Frequency Representations for Testing GSM Equipment," *IEEE Transactions on Instrumentation and Measurement*, October 2000, Vol. 49, No. 5, p. 1050.

Dual-band mobile-phone antenna bends to any shape

Many dual- and multi-frequency antennas have been developed for mobile phones, but four Swiss engineers have developed one that can be bent easily to conform to almost any shape inside the phone's case. J.F. Zurcher, I. Giangrandi, O. Staub, and A.K. Skrivervik of the Ecole Polytechnique Federale de Lausanne designed and built a dual-frequency, single-port printed antenna etched on a thin metalized Kapton film that is conformable to accommodate any shape. The plain inverted F-type antenna (PIFA) incorporates an inductance-capacitance (LC) bandstop filter or "trap" at a selected location to isolate certain portions of the antenna at certain frequencies, enabling dual-band operation. The antenna was designed for the 900- and 1800-MHz Global System for Mobile Communications (GSM) bands. In the 900-MHz band, the antenna achieved a peak gain of −1.9 dBi. In the 1800-MHz band, the antenna achieved a peak gain of −1.5 dBi. See "A Dual-Frequency Printed Conformable Antenna For Mobile Communications," *Microwave and Optical Technology Letters*, December 20, 2000, Vol. 27, No. 6, p. 386.


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A Primer On Using PIN Diodes In VCAs

This tutorial discusses three ways to use PIN diodes to implement a voltage-controlled attenuator for microwave applications.

Part 1 of 2 parts

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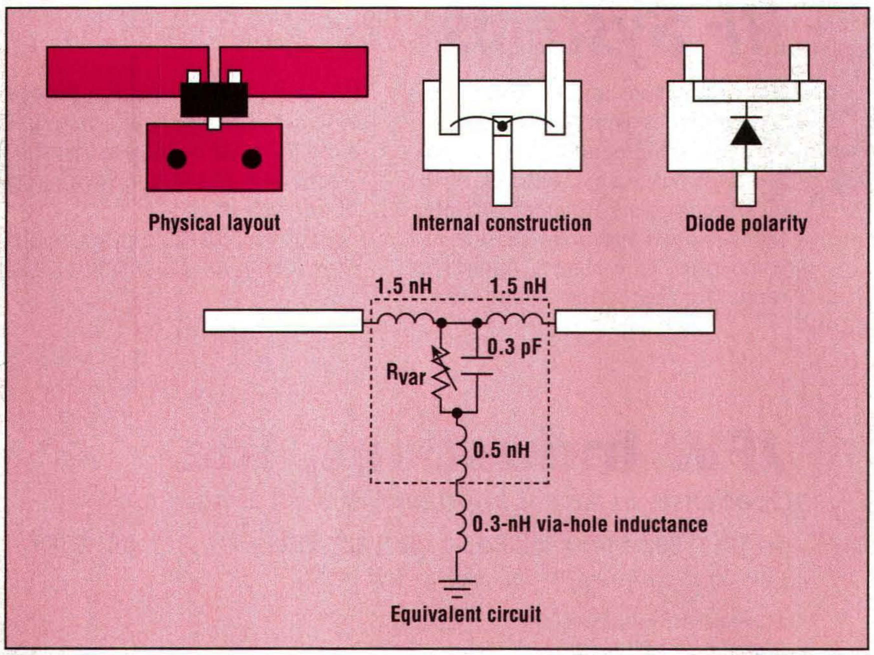
POWER control is an important feature in modern communication systems. In today's spread-spectrum and code-division-multiple-access (CDMA) systems, the signal from each handset appears as thermal noise to the others. If the handset user nearest to the base station produces more power when compared to his neighbor, it will increase the neighbor's noise floor. In the worst case, it will wipe out the other user's signal completely. This is the well-known "near-far" problem. Tight power control can ensure that every user has a quality link to the base station without causing interference to the others. In the receiver (Rx) chain, power control is needed for the automatic-gain-control (AGC) loop to ensure consistency of the signal-to-noise ratio (SNR). Power control also helps to extend the dynamic range of the Rx chain. Power control is typically achieved by using a voltage-controlled attenuator (VCA).

There are many ways to implement a VCA, but it typically involves the use of some variable-impedance device. This variable-impedance device can be a metal-semiconductor field-effect transistor (MESFET) operating as a voltage-variable resistance in its linear region, or it can be a positive-intrinsic-negative (PIN) diode operating as a current-controlled resistance. PIN diodes offer the advantages of high power-handling capability, more design freedom, low distortion, and low cost. This two-part article presents three classic microwave-design approaches using one or more PIN diodes: the resistive-line approach, the constant-impedance approach, and the π-configured, PIN-diode approach.



1. These images show the physical layout, internal construction, diode polarity, and equivalent circuit of the shunt PIN-diode VCA.

PIN-DIODE THEORY

The PIN diode consists of two distinct parts: the die and the package. Figure 1 shows the die modeled as a current-controlled variable resistance and a shunt parasitic-junction capacitance. It consists of a lightly doped I region sandwiched between

heavily doped P-type and N-type regions. The PIN diode behaves as a pure resistance at frequencies 10 times higher than its cutoff frequency f_c, where:

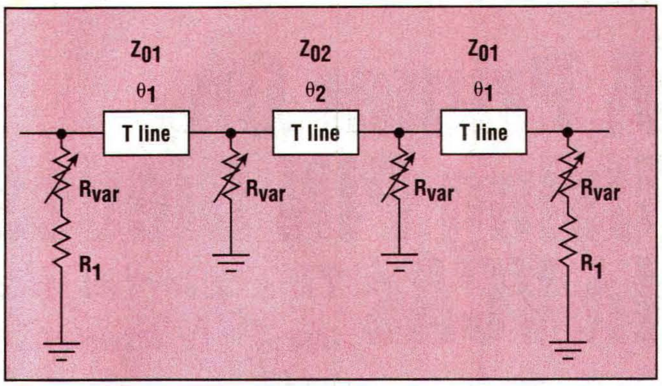
$$f_c = \frac{1}{2\pi\tau} \qquad (1)$$

and τ = the minority-carrier lifetime.

The resistance value of the PIN diode is controlled by DC forward-bias current, which injects carriers into the I-region, lowering its resistance. The PIN diode's resistance can range from a few Ω up to several thousand Ω. I-layer thickness, doping density, and diameter can be adjusted to tailor the diode's characteristics to the specific application. These useful features make the PIN diode an excellent choice for VCA design.

The package that contains the chip adds parasitic inductance and capacitance to the diode's impedance, and today's low-cost leaded plastic packages have particularly large parasitics. In the real world of design, one no longer has the simple variable resistance shown in Fig. 2. However, a simple compensation circuit can be used to tune out the package and chip parasitics at the frequency of interest. For example, a diode package lead and bondwire inductance of 0.7 nH will contribute an inductive reactance of approximately 10 Ω to the diode's resistance. This would seriously degrade the dynamic range of the attenuator, making the compensation circuit very important. The simplest cancellation scheme is to use a shunt capacitor as an RF ground. The shunt capacitor is also used to tune out the parasitic inductance. The value of the shunt capacitor is determined experimentally. The cancellation circuit is used in the resistive-line and constant-impedance approaches. First, consider the resistive-line approach.

The basic building block of the resistive-line approach consists of a



2. This is a schematic of the generic resistive-line approach to VCA design.

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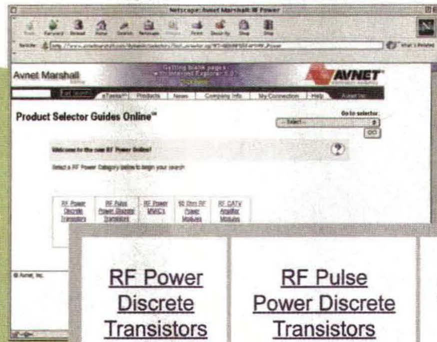

The tool is divided into five categories:

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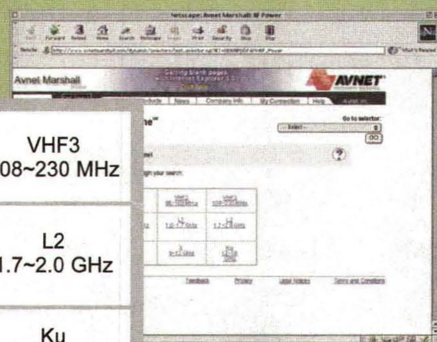

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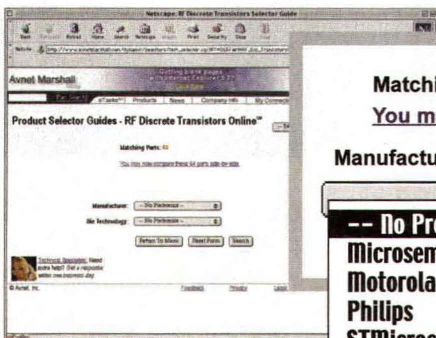
HF 1~30 MHz	VHF1 30~88 MHz	VHF2 88~108 MHz	VHF3 108~230 MHz
UHF1 230~512 MHz	UHF2 512~1000 MHz	L1 1.0~1.7 GHz	L2 1.7~2.0 GHz
S 2.0~4.0 GHz	C 4~8 GHz	X 8~12 GHz	Ku 12~18 GHz

② To narrow the search, specify search parameters within the subheadings of Frequency Range, Die Technology Manufacturer, and Voltage or Application where applicable.



Matching Parts: 64

You may now compare these 64 parts side-by-side.

Manufacturer:	Die Technology:
-- No Preference --	-- No Preference --
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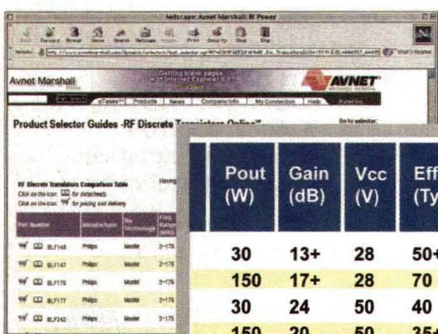
Motorola	LDMOS
Philips	MOSFET
STMicroelectronics	



③ After finding the parts that match your selection criteria, the tool also allows for side-by-side comparisons of similar components:

Pout (W)	Gain (dB)	Vcc (V)	Eff. (Typ)%	RTHj-c (C/W)	Test Freq. (MHz)	Z	Class	Description	Pkg Style	Application
30	13+	28	50+	2.6	175	-	A/B	Comm Source	SOT123	FM Broadcast
150	17+	28	70	0.8	28/108	-	AB/B	Comm Source	SOT121	SSB/FM
30	24	50	40	2.6	28/108	-	A/AB/B	Comm Source	SOT123	SSB/FM
150	20	50	35+	0.8	28/108	-	AB/B	Comm Source	SOT121	SSB/FM
5	16	28	60	11	175	-	A/B	Comm Source	SOT123	FM Broadcast

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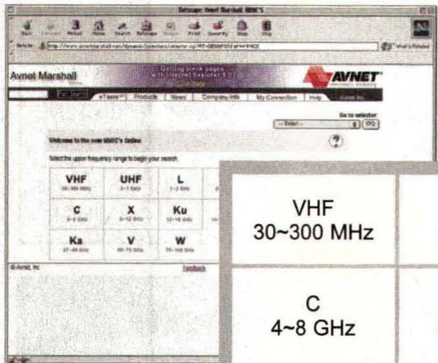

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VHF 30~300 MHz	UHF .3 ~1 GHz	L 1~2 GHz	S 2~4 GHz
C 4~8 GHz	X 8~12 GHz	Ku 12~18 GHz	K 18~27 GHz
Ka 27~40 GHz	V 40~75 GHz	W 75~110 GHz	

① To begin, select the upper frequency range for your search:

Matching Parts: 6

You may now compare these 6 parts side-by-side

Manufacturer:

- -- No Preference --
- Agilent
- EIC
- Infineon
- Maxim
- Motorola
- Philips
- Stanford

Function:

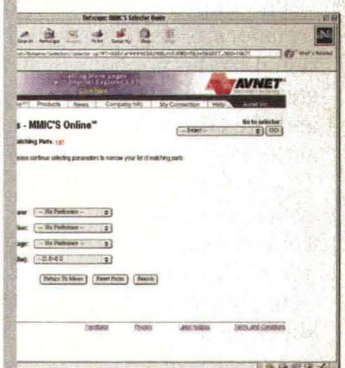
- -- No Preference --
- Gainblock
- LNA
- Power Amp
- Variable Gain Amp

Power (dBm):

- -- No Preference -
- -12.0~0.0
- 0.1~9.0
- 18.0~24.0
- 24.1~37.8
- 9.1~17.9

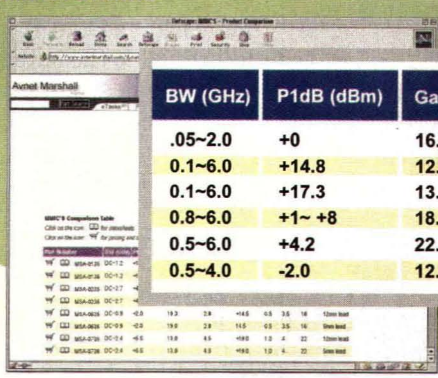
Package:

- -- No Preference --
- 100milstrip
- 145milstrip
- 200milstrip
- 70milstrip
- 85mil SM
- 85milstrip
- SO-8Cer
- SO-8Plstc
- SOT-143
- SOT-25
- SOT-343
- SOT-36
- SOT-363
- SOT-89
- chip



② To narrow the search, provide performance specifications and/or select a specific manufacturer:

③ After finding the parts that match your selection criteria, the tool also allows for side-by-side comparisons of similar components:



BW (GHz)	P1dB (dBm)	Gain (dB)typ	NF (dB)typ	IP3 (dBm)	@ GH	Vd (V)	Id (mA)	Comments
.05~2.0	+0	16.0	2.0	+15.0	0.9	3	1~10	Var Gain
0.1~6.0	+14.8	12.3	2.7	+27.0	2.0	3	42	-
0.1~6.0	+17.3	13.5	2.2	+31.0	2.0	3	84	-
0.8~6.0	+1~ +8	18.5	1.9	+12~ +17	2.0	3	15~50	Var Gain
0.5~6.0	+4.2	22.5	1.6	+15.0	2.4	5	14	-
0.5~4.0	-2.0	12.5	1.6	+8.0	2.4	3	4.5	-

Additionally, the tool also provides access to formal datasheets, online technical support as well as pricing and availability information from Avnet.

please visit: *www.em.avnet.com/rfm*

Table 1: Components used in the resistive-line attenuator

Component	Part No./value	Quantity	Vendor
PIN	HSMP481B	4	Agilent
Resistor	50 Ω	2	KOA
Resistor	110 Ω	1	KOA
Capacitor	10 pF	2	AVX
Capacitor	100 pF	4	AVX

Table 2: Components used in the constant-impedance attenuator

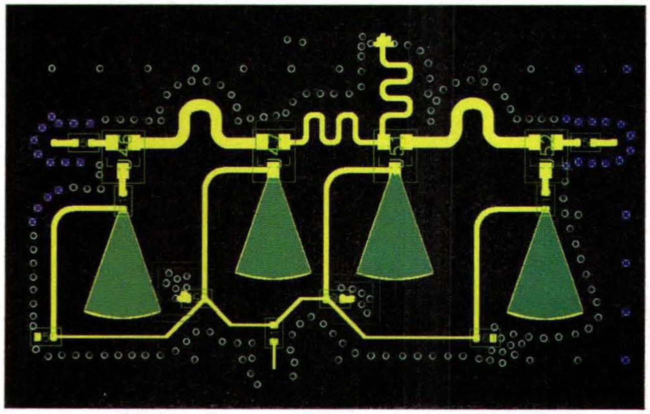
Component	Part No./value	Quantity	Vendor
Hybrid	1A1306-3	1	Anaren
PIN	HSMP3814	1	Agilent
Capacitor	10 pF	5	AVX
Capacitor	1.2 pF	2	AVX
Resistor	910 Ω	2	KOA
Inductor	22 nH	2	Coilcraft

quarter-wave transmission line and a shunt resistance (a PIN diode, in this design). The quarter-wave transformer is a popular way to transform impedance. The design equation is simple:

$$Z_{out} = \frac{Z_0^2}{Z_{in}} \qquad (2)$$

The impedance at one end of the transformer (Z_{in}) is inversely proportional to the impedance at the other end (Z_{out}). For example, if $Z_{in} = 1\ \Omega$ and the transmission line has a characteristic impedance (Z_O) of 80 Ω, Z_{out} will be 6400 Ω. This is a high impedance that will reflect most of the incoming RF signal. Typical values of Z_O for the transmission-line range from 50 to 90 Ω. To obtain variable attenuation using this approach, the fixed value Z_{in} is replaced with variable-resistance PIN diode. At 2.45 GHz, a typical shunt-configured PIN diode can achieve a maximum attenuation of about 20 to 30 dB per stage.

Figure 3 shows a generic resistive-line approach. The particular design discussed here calls for a dynamic range of 50 dB. Therefore, four shunt diodes are used in this design, as shown in Fig. 4. Any number of variable resistances can be used, depending upon the desired range of attenuation. The middle two shunt diodes provide the bulk of the attenuation by reflecting the incoming RF signal. However, a reflective VCA is not desirable in most applications, especially in high-power transmitter (Tx) applications. The reflected RF energy must be absorbed inside the VCA to provide good return loss. This requires the inner resistance to be different from outer resistances. That is why a 50-Ω resistor is used in series with each of the outer PIN diodes.

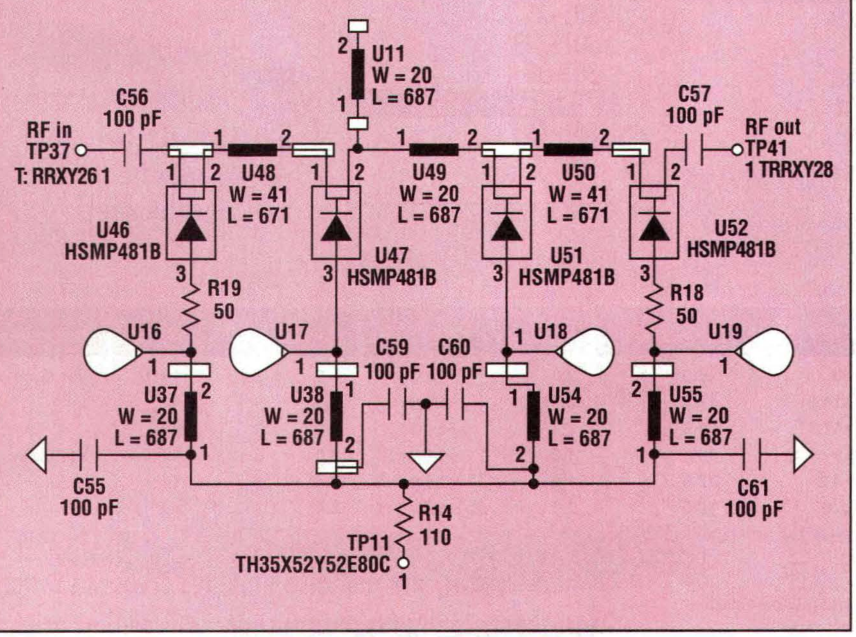


3. This is the circuit-board layout of the resistive-line VCA.

Four identical variable resistors (R_{var}) simplify the PIN diode bias network, and the R_1 fixed resistors are added to the outer variable resistors. The designer can vary the values of R_1, Z_{O1}, Z_{O2}, θ_1 and θ_2 to trade-off size, dynamic range, and input/output (I/O) return loss. In the design discussed here:

$R_1 = 50\ \Omega$
$Z_{O1} = 70\ \Omega$
$Z_{O2} = 95\ \Omega$
$\theta 1 = \theta 2 = 90$ deg.

The circuit shown in Fig. 2 is idealized in that parasitic-diode elements (package inductance and capacitance,



4. The schematic for the resistive-line VCA is shown here.

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CLV0815E	806	824	0.5-4.5	11	-113	-35	5.0	11
CLV0950E	865	1035	1-10	27	-114	-11	5.0	24
CLV0915A	902	928	0-4	17	-108	-30	3.0	10
CLV1085E	1050	1086	0.5-4.5	21	-112	-20	5.0	20
CLV1385E	1370	1400	0.5-4.5	18	-110	-20	5.0	20
CLV1550E	1500	1600	0.5-5.0	44	-106	-35	5.0	22
CLV2465E	2436	2496	1-4	26	-107	-20	5.0	25

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SMV0162A	125	200	0.7-8.3	12	-100	-6	5.0	36
SMV1570L	1540	1600	0.5-2.5	128	-90	-15	2.7	9
SMV2165A	2118	2218	0-3	148	-91	-10	3.3	16
SMV2390L	2290	2485	0-4	116	-90	-11	5.0	16
SMV2660L	2620	2700	0.5-4.5	90	-91	-17	5.0	21

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USSP2330	2300	2360	0.5-2.5	0±3	-83	-15	2.7	8

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PLL0210A	200	230	100	0.50	-105	3.5±2.5	+5	25
PLL0930A	900	960	100	0.75	-101	3±2	+5	40
PLL1260A	1230	1290	1000	0.75	-102	1±2	+5	40
PLL1456A	1420	1490	1000	0.75	-103	1±2	+5	40
PLL2710A	2670	2740	1000	1.25	-98	1±4	+5	30

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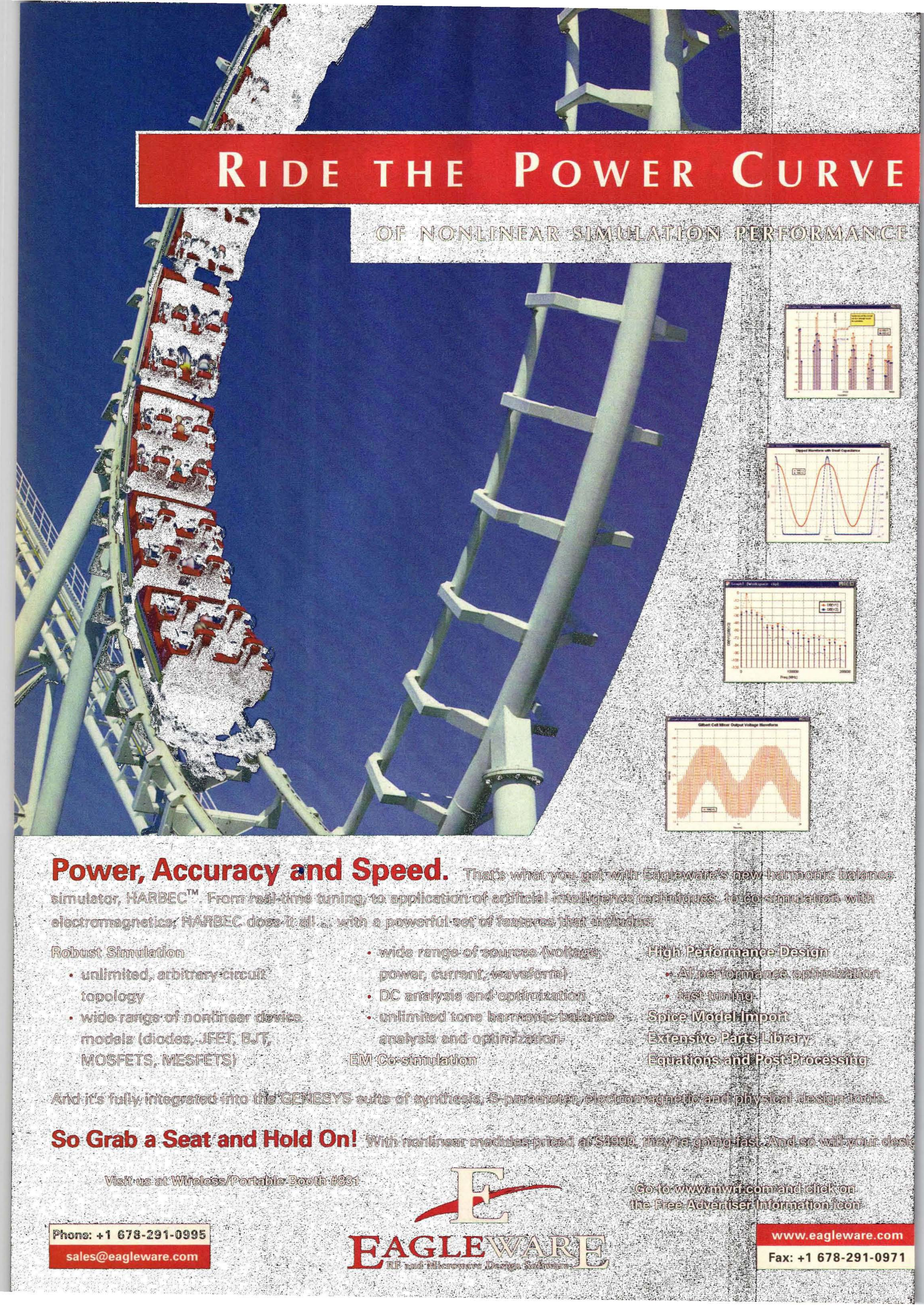
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as well as junction capacitance) are neglected. However, at 2.45 GHz, these elements must be taken into account. The diode chosen for this project is a low-cost SOT-323 part designed for shunt applications, as shown in Fig. 1.

Reducing the effects of package parasitic inductance can turn an ordinary design to a high-performance circuit. The parasitic-inductance cancellation scheme does not have to be complicated. The main contributors to parasitic inductance are package leads, bondwires, and via holes. Each package lead produces 0.5 nH of parasitic inductance, and each bondwire produces 1.0 nH (Fig. 1). Mounting the diode as shown moves the lead and bondwire inductances of leads 1 and 2 into the series circuit, where they cannot reduce the isolation of the diode.

Lead 3 (the GND pin) has 0.5 nH, and via holes (if used) would contribute another 0.3 nH. This inductance can be cancelled by simply using a shortened radial microstrip stub (capacitive impedance) in place of the via holes to resonate out the parasitic inductance of lead 3. A rectangular stub can also be used for this design. The dimensions of the radial stub are determined during simulation.

Having examined the basic building block and parasitic-cancellation scheme, consider the schematic shown in Fig. 4. The circuit is built as a FR-4 microstrip, 0.042 in. (0.106 cm) thick, for lowest cost. The four shunt-configured PIN diodes U46, U47, U51, U52, along with TL U48, U49, and U50, form the skeleton of this design. The radial stubs U16 to U19 are parts of the compensation circuit. U37, U38, U54, and U55 are the RF-choke high-impedance TL. R14 is bias resistance. The other components are capacitors for either DC blocking or bypassing. The circuit has a low component count since free microstrip elements are used as much as possible. Table 1 lists the components.

Figure 3 shows the layout of the VCA, and Figure 5 shows the results of the performance measurements. The major parameters tested are dynamic range and impedance matching at each attenuation state. The prototype demonstrated approximately 50-dB dynamic range. The input and output impedance match was better than –10 dB in all attenuation states. The current consumption was only 48.3 mA for maximum attenuation.

Figure 6 shows a generic design approach. A quadrature (90-deg.) coupler and two variable resistances can be used to form a low-cost variable attenuator having moderate dynamic range and good input and output impedance matching. An RF signal is applied to port 1 of the coupler and two matched variable resistances on ports 2 and 3 are varied in magnitude. The result is a variable

attenuation between ports 1 and 4. The input impedance at port 1 and the output impedance at port 4 remain constant at 50 Ω over the entire range of attenuation. This characteristic is assured by the fundamental operation of a quadrature (Q) hybrid.

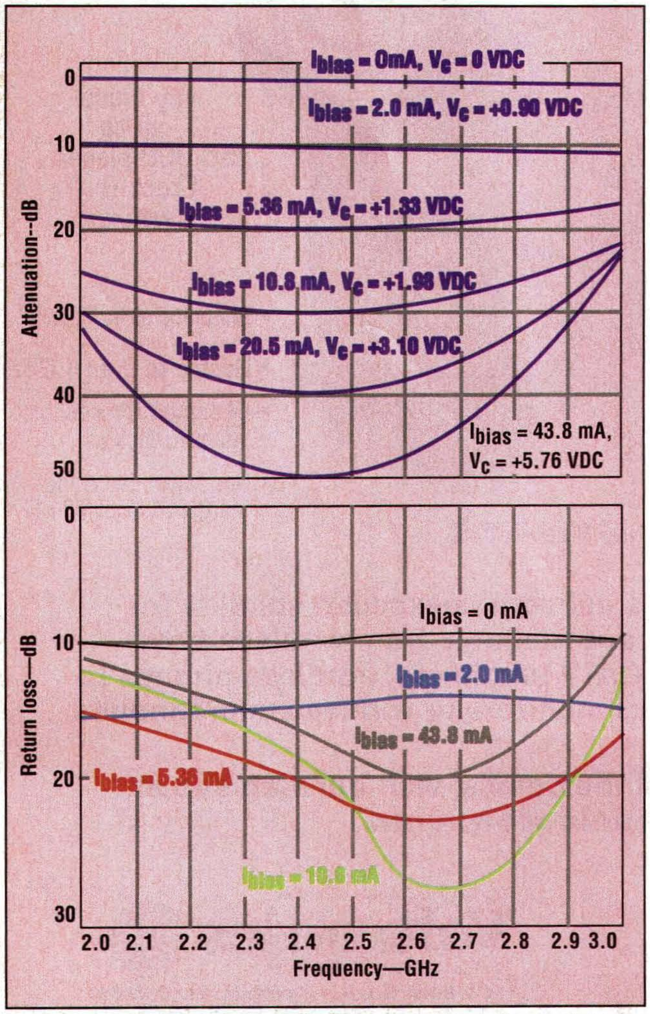
DIODE RESISTANCE

When the PIN-diode current is high (>10 mA), diode resistance is low and the RF signal applied at port 1 is reflected back into the hybrid at ports 2 and 3, emerging at port 4 with little loss. As current is reduced, diode resistance rises to a value of 50 Ω, at which point the diodes absorb the incident RF signal and attenuation is highest. Further reduction in current increases diode resistance to values higher than 50 Ω, resulting in reflections at ports 2 and 3 and lower attenuation. Figure 7 shows attenuation versus forward current for a typical PIN diode with a thick I-layer. The designer clearly has two current-range options: 0 to 1.7 mA, or 1.7 to 100 mA. Each choice has its trade-offs. If the current range I > 1.7 mA is used, some circuit simplification may be possible. But minimum attenuation remains high (on the order of 2 dB or more), due to the fact that the PIN-diode resistance is not zero at high currents. If the current range I <1.7mA is used, circuit insertion loss (minimum attenuation) will be lower. This article discusses the latter approach to the design.

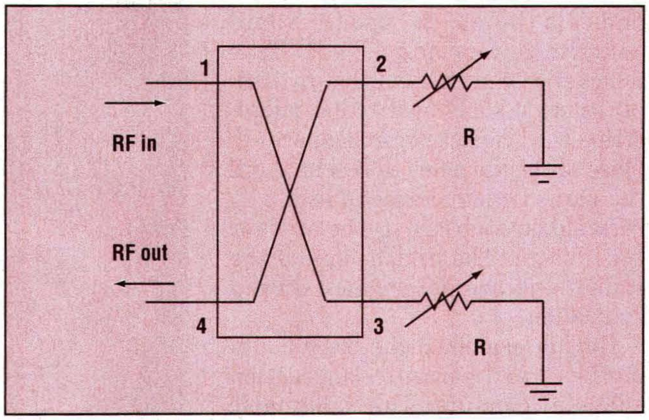
Figure 8 shows the microstrip board layout. The circuit was built on 0.014-in. (0.036-cm)-thick FR4 microstrip with half-oz. Copper (Cu) conductor and ground plane.

A few simple experiments were performed on the hybrid evaluation board to verify the concept at the two extremes of diode resistance. In the first case, ports 2 and 3 were left open-circuited (no diode), yielding an insertion loss of 1.5 dB from port 1to port 4. In the second case, ports 2 and 3 were terminated with a 50-Ω load (chip resistance). The resulting value of S41 was more than 30 dB. Table 2 lists the parts used in the evaluation board, and Figures 9 and 10 show the results of the evaluation. Figure 11 shows the schematic for the attenuator.

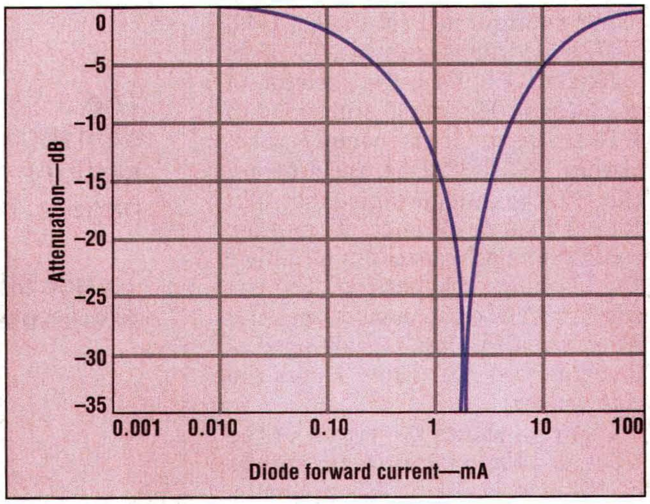
An observant reader may wonder why there is a deep notch at the maximum attenuation. The answer lies in the parasitic-compensation circuit. At the series resonant frequency, the parasitic inductor and compensation capacitor cancelled each other out. The Q of the resonant frequency is



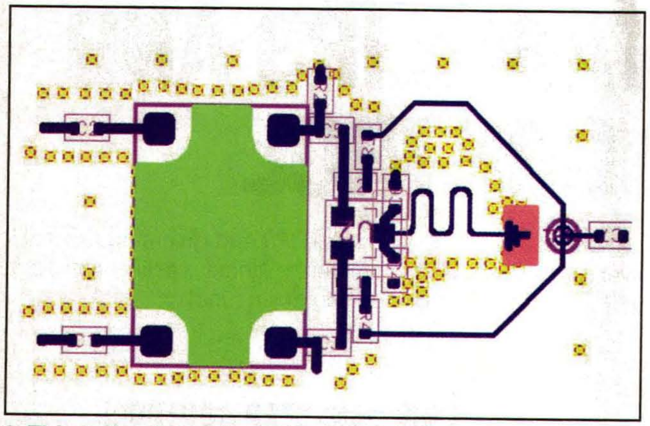
5. These graphs plot the data measured from the resistive-line VCA.



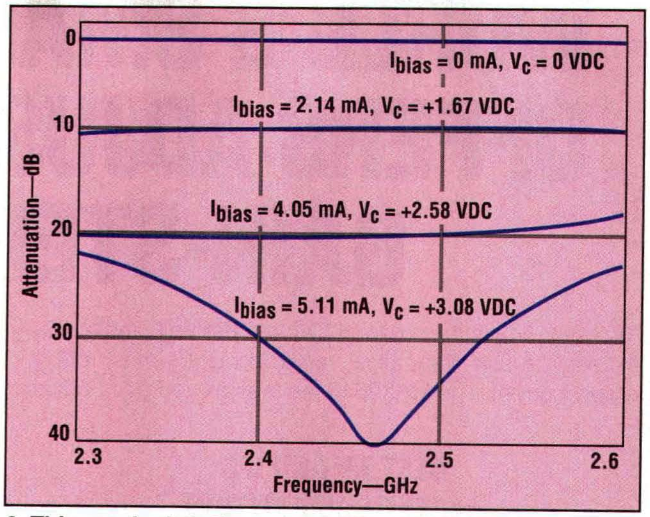
6. This is the schematic of the generic constant-impedance approach to VCA design.



7. This graph plots the PIN diode's attenuation versus forward current in the constant-impedance VCA.



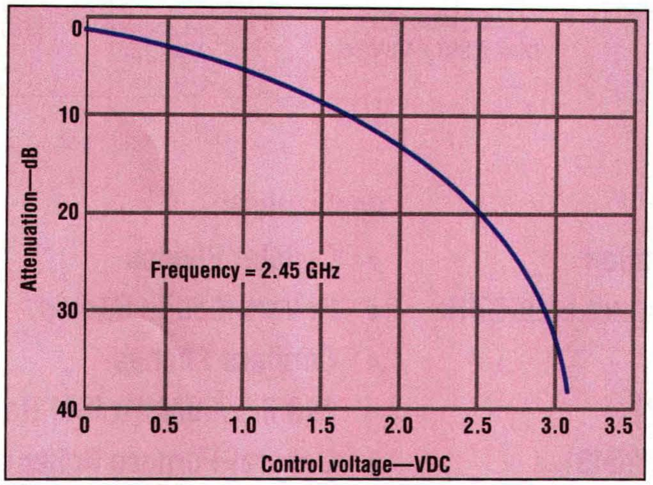
8. **This is the circuit-board layout of the constant-impedance VCA.**



9. **This graph plots the constant-impedance VCA's attenuation versus frequency.**

high. That is what causes the deep notch. The VCA easily achieves 30-dB dynamic range and offers more control range near the center of the frequency band.

Further improvement or modification can be made to make the design either more compact or low cost. This design used an off-the-shelf, surface-mount hybrid coupler. A hybrid based on another technology, such as lumped elements, may be smaller, making the VCA more compact. If printed-circuit-board (PCB) real estate is not a problem, a transmission-line coupler design such as the popular and high-performance branchline coupler can be used. In this case, the coupler is "free" because it is part of the etched microstrip board. The RF choke inductor can also be replaced with high-impedance transmission lines to further reduce the cost if there is enough board space available.



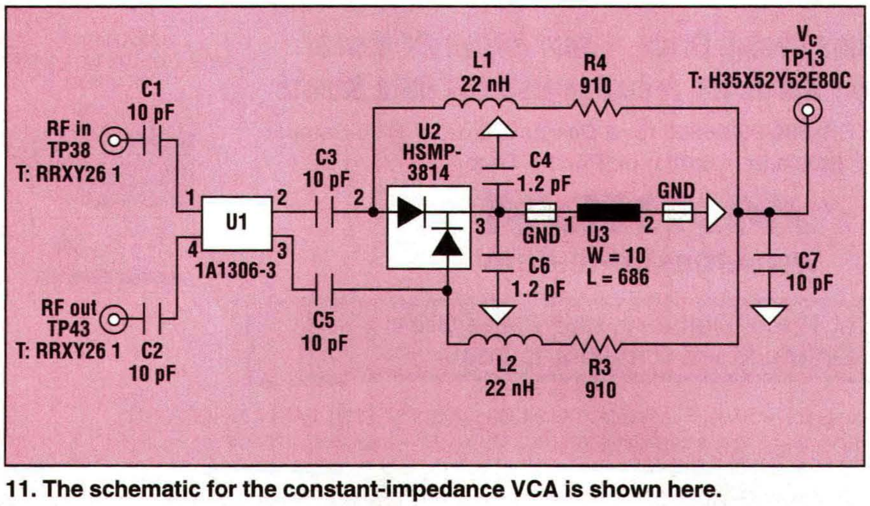
10. **This graph plots the constant-impedance VCA's attenuation versus control voltage.**

Part 2 of this article will cover the π-configured VCA and discuss the advantages and disadvantages of all three design approaches.

APPENDIX: THE 3-dB QUADRATURE COUPLER

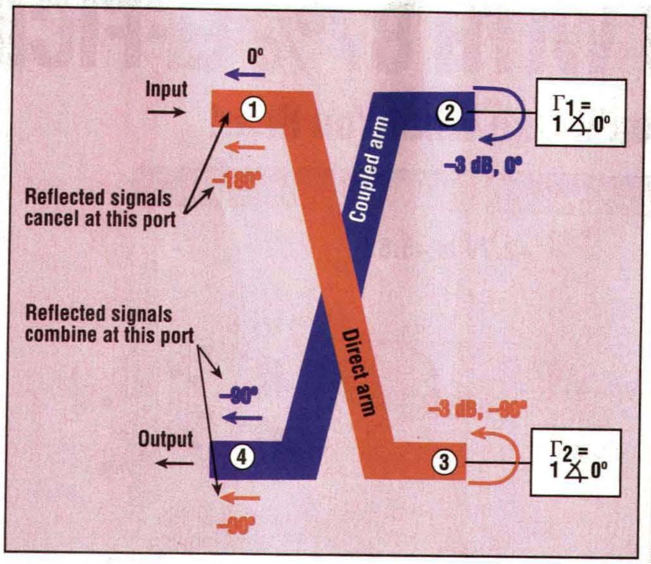
The 3-dB quadrature coupler can take three general forms: as a distributed-transmission-line, backward-wave coupler of one or more sections (as illustrated in Figs. 12 and 13), as a transmission-line branchline coupler, or as a lumped-element device. All three forms share some common characteristics, which make this RF component particularly useful. If an RF signal is applied to port 1 (known as the "input"), half the power will come out at port 3 (called "direct") and half will come out at port 2 (known as "coupled"), and the equal-amplitude RF voltages at these two ports will differ in their phase angles by 90 deg. In the ideal coupler, nothing will come out of port 4, known as the "isolated" port.

However, if one places mismatches of equal magnitude and phase angle on the outputs 2 and 3, some of the RF energy will be reflected back into the quadrature coupler. In the simple case shown in Fig. 12, ($\Gamma_1 = \Gamma_2 = 1.0$ /0), the RF energy will recombine at the "isolated" port (minus 2× the insertion loss of the coupler). If $\Gamma_1 = \Gamma_2$ with a magnitude less than 1.0, the attenuation between the "input" and

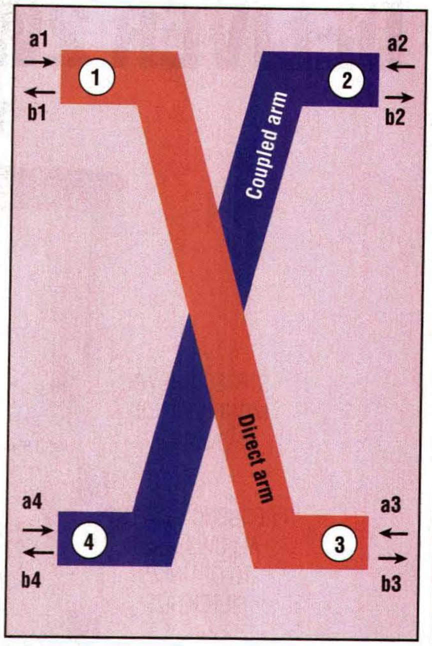


11. **The schematic for the constant-impedance VCA is shown here.**

"isolated" ports will be 20 $\log_{10} \Gamma$ (the return loss of the loads) plus 2× the insertion loss of the coupler. If one can realize a current or voltage-controlled variable mismatch ($0 > \Gamma > 1$), a simple variable attenuator can be created. When the package parastics and junction capacitance of the PIN diode are tuned out with a simple single capacitor or inductor, the diode's junction can be made to vary from 2 to 3000 Ω, passing through 50 Ω.The foregoing heuristic analysis may not satisfy those looking for a more rigorous analysis. Figure 13 defines the four-port S parameters. The familiar S-parameter definitions, such as:



12. This is the general form of a –3-dB quadrature coupler.



13. This quadrature-coupler diagram defines the four-port S-parameters.

$$S11 = \frac{b1}{a1} \qquad (1)$$

$$S32 = \frac{b3}{a2} \qquad (2)$$

apply. One can write the equation for the operation of the quadrature coupler as:

$$[b1\ b2\ b3\ b4] = \begin{bmatrix} S11 & S12 & S13 & S14 \\ S21 & S22 & S23 & S24 \\ S31 & S32 & S33 & S34 \\ S41 & S42 & S43 & S44 \end{bmatrix} \times \begin{bmatrix} a1 \\ a2 \\ a3 \\ a4 \end{bmatrix} \qquad (3)$$

For the ideal (zero loss, infinite isolation) 3-dB quadrature coupler, the S-parameters are:

$$S11 = S22 = S23 = S32 = S33 = S44 = 0 \qquad (4)$$

$$S12 = S21 = S34 = S43 = 1 + j0 \qquad (5)$$

$$S13 = S24 = S31 = S42 = 0 + j1 \qquad (6)$$

If a1 is set to 1 and a2 = a3 = a4 = 0, multiplying the matrices will result in the solution:

$$\left[0 \quad \frac{-1}{\sqrt{2}} \quad \frac{-j}{\sqrt{2}} \quad 0\right] \qquad (7)$$

in which it can seen that the magnitude of b2 and b3 is 0.707, and that they differ in phase by 90 deg.

Now, consider a mismatch on ports 2 and 3 of $\Gamma_2 = \Gamma_3 = \rho/0°$, where:

ρ = the magnitude of the reflection coefficient. Referring to Fig. 8 in the main article,

$$a3 = \rho\frac{-j}{\sqrt{2}} \qquad (8)$$

and

$$a2 = \rho\frac{-j}{\sqrt{2}} \qquad (9)$$

Plugging these values into the equation for the quadrature coupler, one obtains the solution:

$$[0\ \ 0\ \ 0\ \ \rho] \qquad (10)$$

or S41 for the entire network is ρ, resulting in an insertion loss of 20 $\log_{10} \rho$. This is the same result as that obtained in our intuitive explanation.

The foregoing discussion assumes that the coupler has infinite isolation (loss from "input" to "isolated" with perfect 50-Ω terminations on "coupled" and "direct" outputs). In practice, no coupler has perfect isolation, and the designers must choose their couplers carefully to ensure that they offer greater isolation than the dynamic range demanded of their attenuators. ••

Acknowledgement

The authors thank Mauren Bennett of Agilent Technologies, Peter Shveshkeye, Gerald Hiller, Todd Brown of Alpha Industries, and Chad Blitz of Anaren for their time, assistance, and generous discussions on technical issues.

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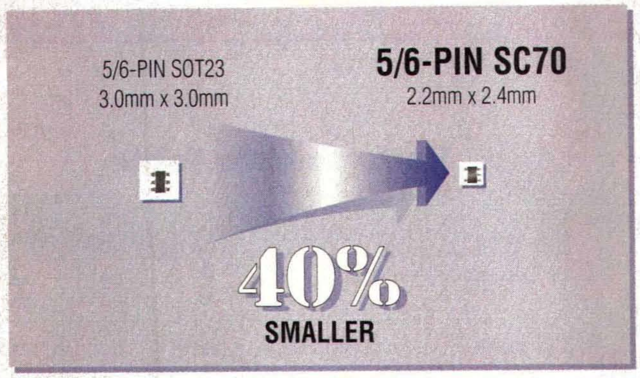

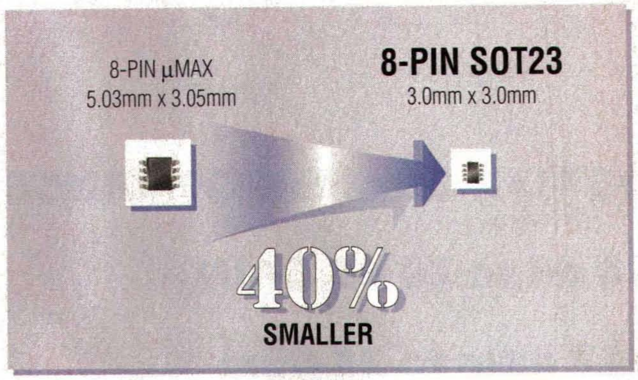

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EDA Tool Relates EVM To A Filter's Group Delay

A design automation tool supports the calculation of error-vector magnitude (EVM) from a filter's group-delay parameters.

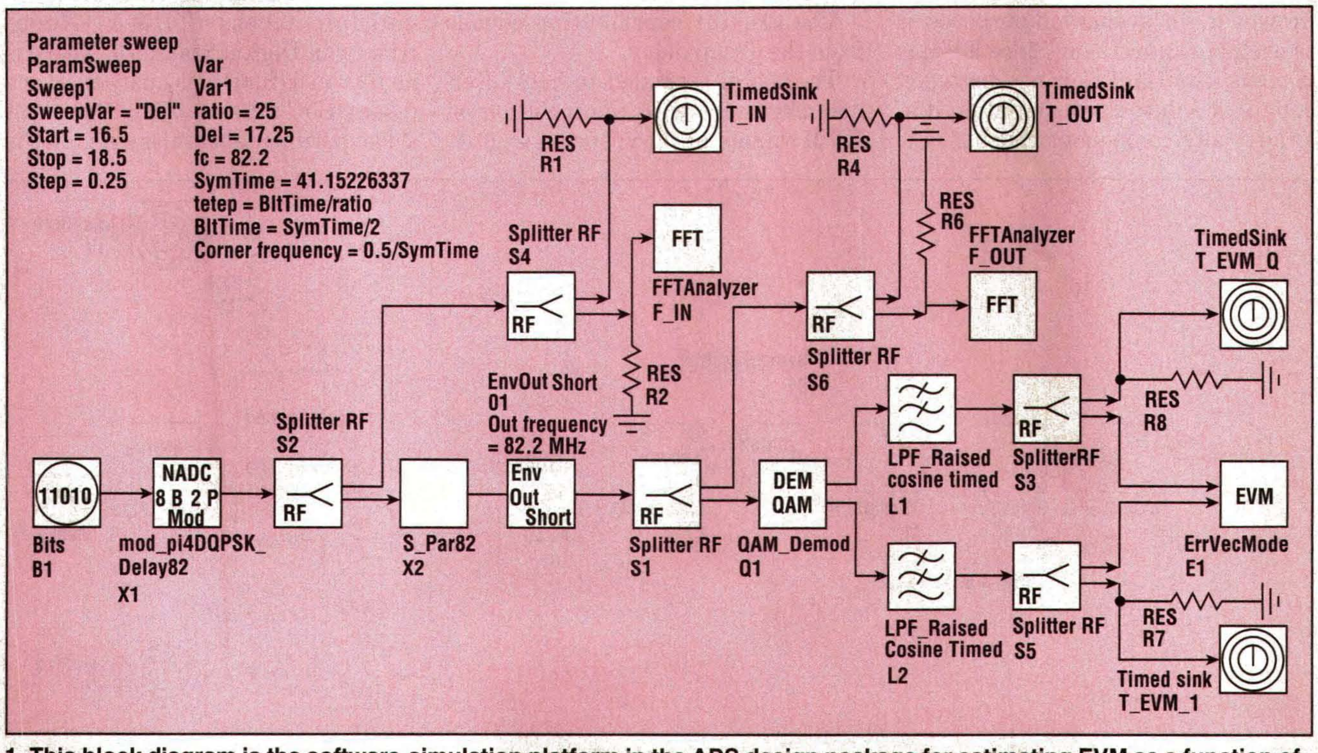
Boris Aleiner

Staff Engineer

Motorola, Inc., 330 S. Randolphville Rd., Piscataway, NJ 08854; (732) 743-7489, FAX: (732) 878-8124, e-mail: Boris_Aleiner-W15279@email.mot.com.

MODERN digital transmitters (Txs) in wireless communications systems must meet a specific value of error-vector magnitude (EVM) which defines the modulation accuracy of the system. EVM is a measure of the difference between a Tx's input- and output-constellation diagrams. This permits some estimate for the value of bit-error rate (BER), a receiver (Rx) characterization that determines the quality of digital communications.

It turns out that a filter's group-delay value has an influence on the value of EVM. In this article, an electronic-design-automation (EDA) tool will be shown to enable the prediction of the group delay's influence on EVM value. To do this, a software-simulation platform based on the Agilent Technologies advanced-design-system (ADS) EDA package has been developed. Simulations of a measured S-parameter filter file are



1. This block diagram is the software-simulation platform in the ADS design package for estimating EVM as a function of group delay. The basic elements are a signal-source input (X1) and a generic RF filter (X2).

conducted, and the results of these simulations are confirmed experimentally by measurements on a physical bandpass filter. The simulated and measured results coincide very closely.

The formula used for EVM calculations is provided by,[1]

$$V_{mod}(t) = A\begin{bmatrix} I_{enc}(t)\cos(\omega_c t) - \\ \Delta A Q_{enc}(t)\sin(\omega_c t + \varphi) \end{bmatrix} \quad (3)$$

where:

MIN and MAX are the first and the last symbols within the symbol burst for which EVM is to be measured, and

E(k) is the error vector, calculated as a difference between the actual and the expected position of the signal vector on the constellation diagram.

To obtain meaningful results for EVM calculations, Eq. 1 has to be used within a software package such as the ADS program.

As shown by Eq. 1, it is difficult to determine precisely the individual contributions of a Tx's component chain to the total Tx EVM value. However, for evaluation purposes, it is useful to know them. Since EVM is a measure of a difference between input- and output-constellation diagrams, any component causing output-constellation degradation contributes to the EVM degradation. The components that degrade the output-constellation diagram are nonlinear components (amplifiers and mixers) and components causing signal delays (filters).

There are two types of delay that the modulated signal-vector experiences in the Tx path. One of them is phase delay, a measure of the modulated signal-vector's phase change. The second is group delay, a measure of the modulated signal vector's amplitude change. The formula describing a signal vector at the output of a Tx path is provided by, [2]

$$y(t) = Kx_c\left(t - \tau_g\right) \times \cos\left[2\pi f_c\left(t - \tau_p\right)\right] \quad (2)$$

where:

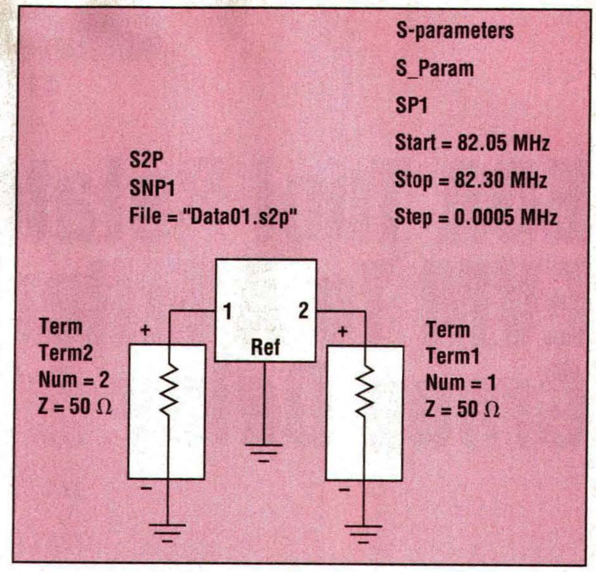
τ_g is the group delay, and

τ_p is the phase delay.

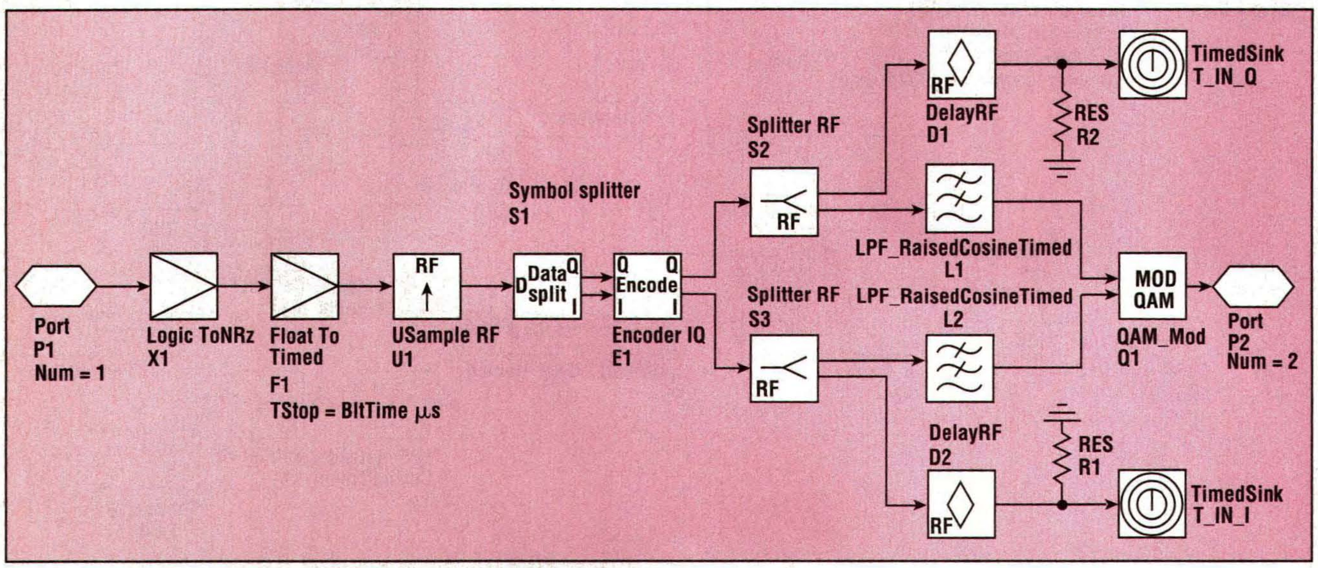
The greater contribution comes from the group delay.

The tool is intended to help filter manufacturers with the inclusion of EVM values in filter specifications, as is sometimes requested by customers. This would make it possible to supply EVM values of filters without the need to conduct measurements later on expensive equipment. This tool can also be useful for system architects in evaluations of Tx-component selections.

This article will focus on the $\pi/4$ differential-quadrature-phase-shift-keying (π/4DQPSK) modulation scheme used for the time-division-multiple-access (TDMA) North American Digital Standard. There is no reason to limit this analysis to one modulation scheme other than that the experimental results support the



3. The generic filter model (X2 in Fig. 1) in ADS is an S-parameter file implemented with the software's RF/Analog bench utility. The S-parameters determine the filter's group delay.



2. The I and Q signals required to generate an input constellation diagram are generated by this ADS simulation. This simulation is the model of the signal source X1 in Fig. 1.

simulations performed. There are simulation platforms available for all existing modulation schemes. While all would provide correct results, the only one with proven results is π/4DQPSK modulation scheme.

THE SIMULATION

To calculate EVM values as a function of a group delay, a simulation platform has been developed (Fig. 1). It is based on one of the platforms reported earlier and consists of two basic elements.[3]

The first element, labeled X1 in Figs. 1 and 2, is a signal source. The input random-bit stream is converted into pulses by the float-to-timed converter (F1 in Fig. 2) and differentially encoded (according to IS-136 standard requirements) by in-phase/quadrature (I/Q) encoder E. It is then shaped by the root-raised cosine filters (L1 and L2), and RF modulated by the quadrature-amplitude-modulation (QAM) modulator Q1. RF modulation is performed according to the formula:

$$V_{mod}(t) = A\begin{bmatrix} I_{enc}(t)\cos(\omega_c t) - \\ \Delta A Q_{enc}(t)\sin(\omega_c t + \varphi) \end{bmatrix} \quad (3)$$

where:

$I_{enc}(t)$ and $Q_{enc}(t)$ are the differentially encoded I and Q bit streams, respectively,

ΔA represents an amplitude imbalance between I and Q bit streams, and

π represents a phase imbalance in the I and Q bit streams.

The I and Q bit streams are measured by so-called time sinks, T_IN_I and T_IN_Q, which are components for measurements of signals (Fig. 2). They are used to determine an input-constellation diagram.

The second element of a simulation platform, labeled X2 (Fig. 3), is a generic RF filter. This element is implemented in the RF/Analog bench of ADS. That tool allows a designer to use component libraries or to conduct evaluations using other RF methods (such as S-parameters). The value of a filter's group delay is calculated by that bench based on the

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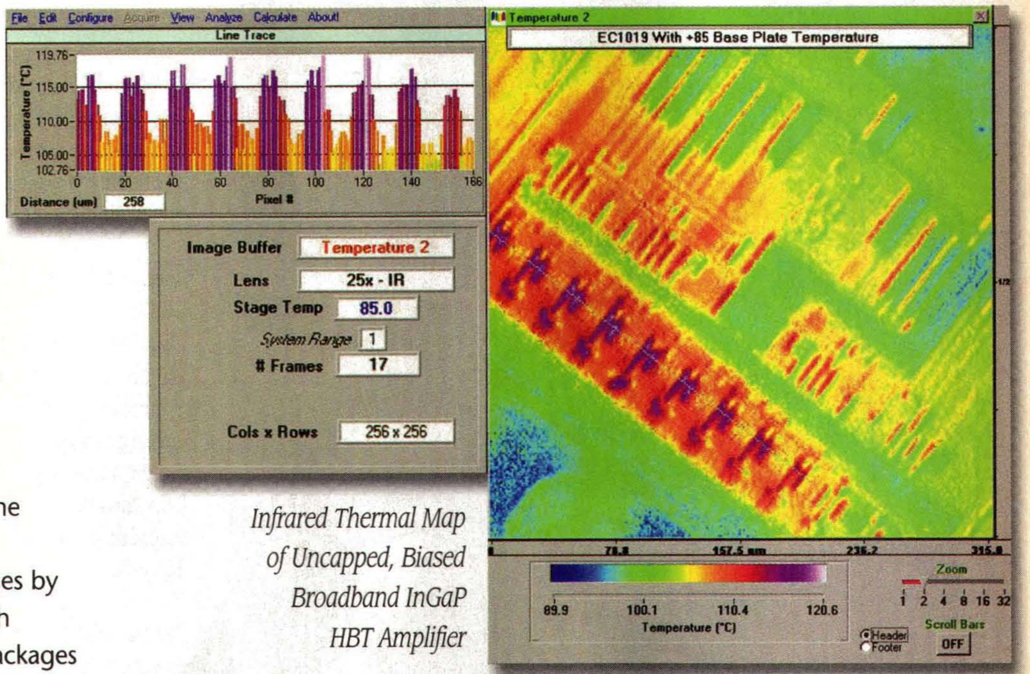

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ECG002	20dB	15dBm	29dBm	233° C/W	40°C	DC-6 GHz
ECG006	15dB	15dBm	30dBm	278° C/W	50°C	DC-6 GHz
ECG003	20dB	23dBm	39dBm	50° C/W	45°C	DC-3 GHz
ECG008	15dB	23dBm	40dBm	55° C/W	55°C	DC-3 GHz
ECG009	19dB	24dBm	41dBm	85° C/W	65°C	DC-2 GHz
ECG011	20dB	8dBm	20dBm	355° C/W	47°C	DC-6 GHz
ECG012	14dB	20dBm	36dBm	120° C/W	45°C	DC-2.5 GHz
EC-1089	15dB	23.5dBm	>42dBm	~85°C/W	~65°C	DC-2.5 GHz
EC-1019	18.5dB	19dBm	34dBm	120°C/W	40°C	DC - 3 GHz
EC-1078	19.5dB	21dBm	37dBm	120°C/W	60°C	DC - 3 GHz
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Calculate Oscillator Jitter By Using Phase-Noise Analysis

Two types of jitter specifications can be determined by developing equations based on analyzing an oscillator's phase noise.

Part 1 of 2 Parts

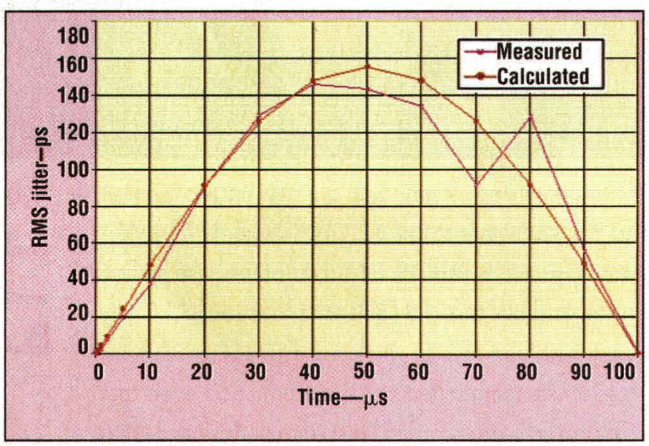
Boris Drakhlis
SaRonix, 141 Jefferson Dr., Menlo Park, CA 94025-1114; (800) 227-8974, (650) 470-7700.

OVER the last several years, jitter has become a significant and important parameter for characterizing short-term stability of crystal oscillators in the time domain. This is driven by applications that use crystal oscillators as clock sources in high-performance computer, networking, and communications equipment. Traditionally, Allan variance has been used for characterization of short-term frequency stability for crystal oscillators in the time domain. The relation between Allan variance (time domain) and phase noise (frequency domain) is described in numerous papers.[1]

Theoretically, jitter is defined as short-term non-cumulative variations of the significant instants of a digital signal from their ideal positions in time.[2] In practice, two methods of measuring jitter are used: period-jitter measurements with a digital storage oscilloscope (DSO) or time-interval analyzer (TIA), and phase-jitter measurements in a specified frequency band recalculated to the time domain. The last method is used, for example, in Synchronous Optical Network (SONET) equipment-jitter specifications.

Of importance is the fact that there is a distinct difference between the period and phase methods of measuring jitter. Therefore, it is impossible to determine whether an 155.52-MHz clock oscillator with 1-ps period jitter measured with a DSO will work in a SONET application that requires 155.52-MHz clock frequency with the maximum reference clock jitter in a 12-kHz-to-20-MHz band of 1-ps root mean square (RMS).

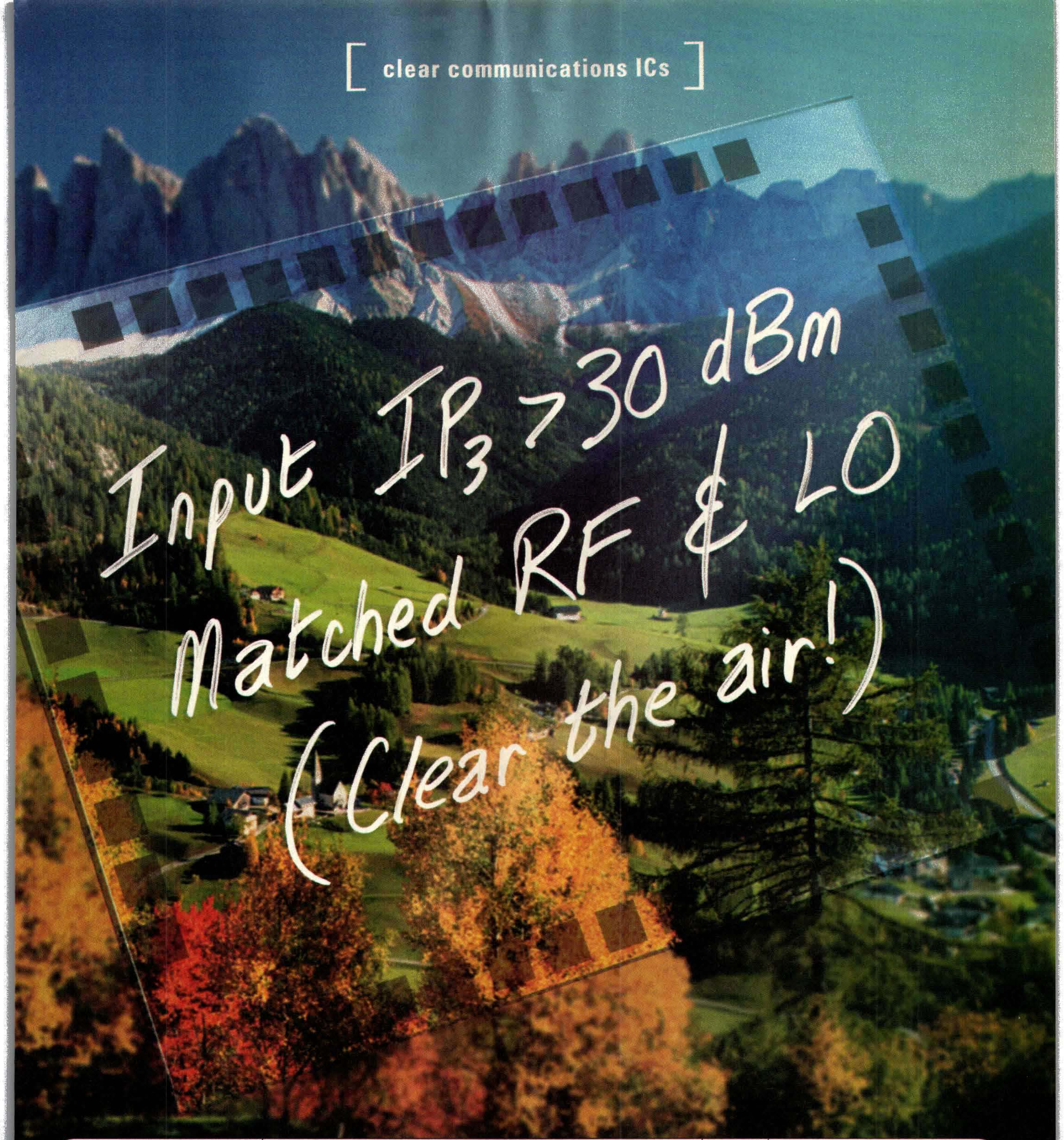
This article illustrates how the phase noise of an oscillator can be used to calculate the period jitter and the RMS jitter in the specified band. The equations developed will be applied to several practical examples, including crystal-clock oscillators and phase-locked loops (PLLs).



1. These plots of measured and calculated modulation-induced jitter for a single level of modulation at –43 dB and at 10 kHz show close agreement.

JITTER AND PHASE NOISE

Jitter accumulated for a number of clock periods ("N") is typically defined as the RMS deviation of N periods from the average value.



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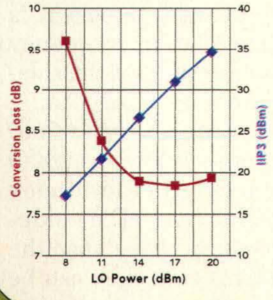

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The one-period jitter that is most frequently used for characterization of clock sources corresponds to N=1.

Consider the well-known model of the oscillator signal with the absence of amplitude modulation:

$$V(t) = V\sin\left(2\pi f_0 t + \varphi(t)\right) \quad (1)$$

where:

f_0 is the oscillator nominal frequency, and $\phi(t)$ is the oscillator phase noise.

Jitter measurements consist of measuring the time between zero crossings of Eq. 1. In the case of measuring jitter accumulated for N periods, we have the following set of equations:

$$V(t_1) = 0 \quad (2a)$$

$$V(t_2) = 0 \quad (2b)$$

From this and Eq. 1:

$$2\pi f_0 t_1 + \varphi(t_1) = 0 \quad (3a)$$

$$2\pi f_0 t_2 + \varphi(t_2) = 2\pi N \quad (3b)$$

Subtracting the first equation from the second:

$$2\pi f_0(t_2 - t_1) + \varphi(t_2) - \varphi(t_1) = 2\pi N \quad (4)$$

By definition:

$$t_2 - t_1 = NT_0 + \Delta t \quad (5)$$

where:

$T_0 = 1/f_0$ and Δt is the jitter accumulated for N periods.

Substituting 5 in 4:

$$2\pi \frac{1}{T_0}(NT_0 + \Delta t) + \varphi(t_2) - \varphi(t_1) = 2\pi N \quad (6)$$

or

$$2\pi N + 2\pi \frac{\Delta t}{T_0} + \varphi(t_2) - \varphi(t_1) = 2\pi N \quad (7)$$

The $2\pi N$ terms cancel out and after rearranging the remaining terms:

$$\Delta t = \frac{T_0}{2\pi}\left(\varphi(t_1) - \varphi(t_2)\right) \quad (8)$$

In this equation, $\phi(t_1)$ and $\phi(t_2)$ are random functions of time, and Δt is a statistical quantity. To obtain the RMS value of Δt square Eq. 8 and average the result:

$$\left\langle \Delta t^2 \right\rangle = \frac{T_0^2}{4\pi^2} \times \left(\left\langle \varphi(t_1)^2 \right\rangle - 2\left\langle \varphi(t_1) \times \varphi(t_2) \right\rangle + \left\langle \varphi(t_2)^2 \right\rangle \right) \quad (9)$$

Here $\phi(t)$ is a stationary process, and:

$$\left\langle \varphi(t_1)^2 \right\rangle = \left\langle \varphi(t_2)^2 \right\rangle = \left\langle \varphi(t)^2 \right\rangle = \int_0^\infty S_\varphi(f)df \quad (10)$$

where:

S_ϕ is the spectral density of $\phi(f)$, and f is the Fourier frequency.[1,3] Also:

$$\left\langle \varphi(t_1) \times \varphi(t_2) \right\rangle = R_\varphi(t_2 - t_1) = R_\varphi(\tau) = \int_0^\infty S_\varphi(f)\cos(2\pi f\tau)df \quad (11)$$

where:

$R\phi(\tau)$ is the autocorrelation function of $\phi(f)$ and $\tau = t_2 - t_1 \approx NT_0$ in the case of jitter measurements.

Substituting 10 and 11 in 9 gives:

$$\Delta t_{RMS}^2 = 2\frac{T_0^2}{4\pi^2}\int_0^\infty S_\varphi(f) \left(1 - \cos(2\pi f\tau)\right)df = 2\frac{T_0^2}{4\pi^2}\int_0^\infty S_\varphi(f) \times 2\sin^2(\pi f\tau)df \quad (12)$$

or

$$\Delta t_{RMS}^2 = \frac{T_0^2}{\pi^2}\int_0^\infty S_\varphi(f) \times \sin^2(\pi f\tau)df \quad (13)$$

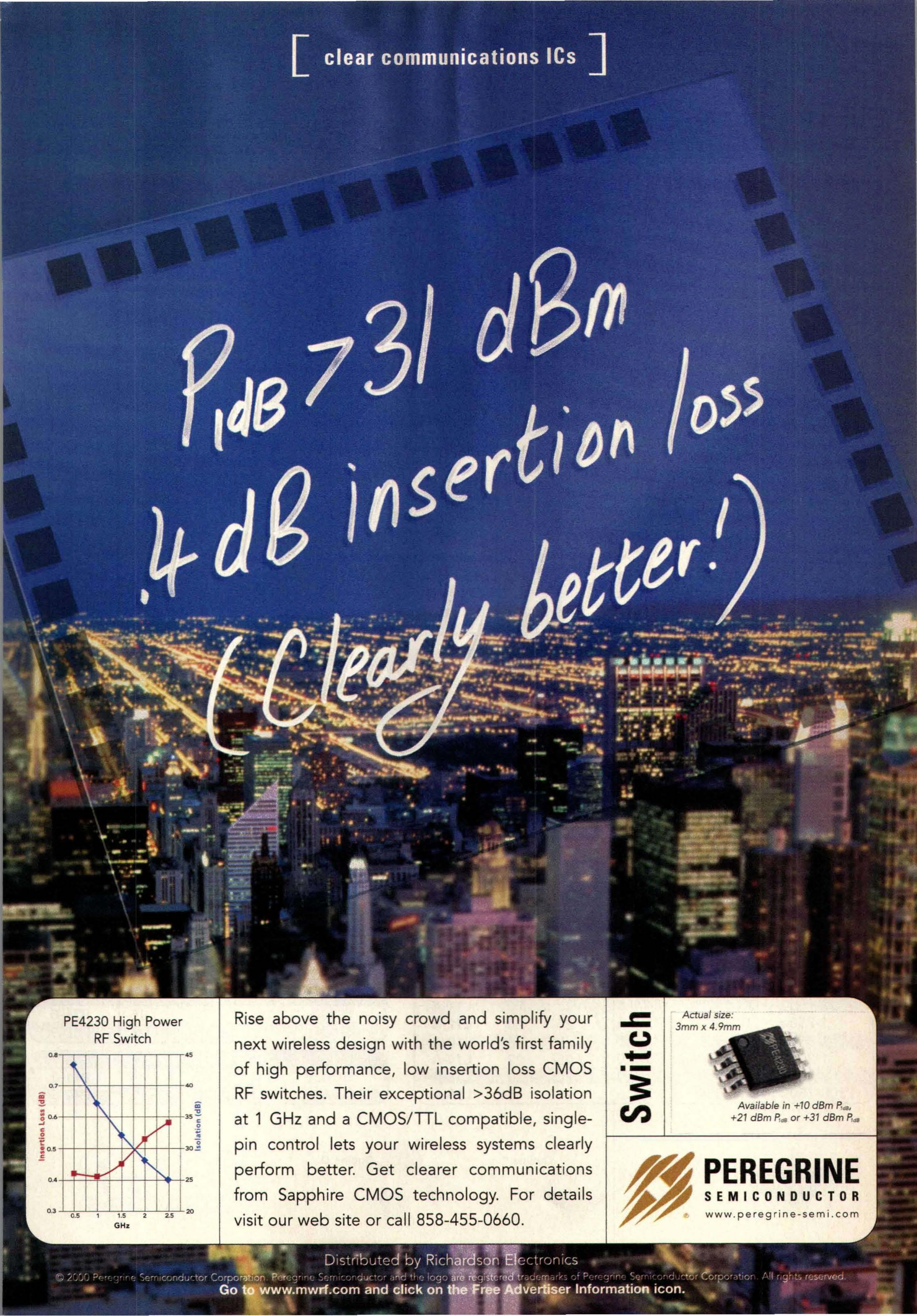
This equation shows that with respect to the period-jitter generation, the phase noise of a signal is filtered by a function that depends on the Fourier frequency and the time between measurements.

The integral in Eq. 13 is taken from 0 to ∞. In practice, the high-frequency cutoff f_h is always present either in the device being measured or in the measuring equipment itself.[3] The measurement time defines the low limit in Eq. 13.

It does not seem as though there should be any problem in confirming the results obtained in these calculations. All that is necessary is to measure the oscillator phase noise, calculate jitter, and compare it to the period jitter that is directly measured with, for example, the DSO. But in reality, the problem is more complicated. In an actual oscillator, amplitude noise exists along with phase noise. The amplitude noise is converted to phase noise on the input of the DSO[4] and thus contributes to the overall jitter reading. The jitter is measured relative to the time base of the DSO, which also has a jitter characteristic. That means that the measured value depends upon the jitter of the DSO and the oscillator under test. If the DSO's timebase jitter is higher than the oscillator jitter, it is the DSO's jitter rather than the oscillator jitter that will be measured. Therefore, in order to substantiate Eq. 13, we need some kind of reference source with a known jitter. A voltage-controlled crystal oscillator (VCXO) that has been modulated at a known frequency and level could be used as such a source.

FREQUENCY MODULATION

It is known that for small modulation indexes, frequency modulation produces two sidebands positioned symmetrically, relative to the carrier. These sidebands create a single linear spectral component in $S_\phi(f)$. The frequency of this component is equal to the modulation frequency and its level is twice that of the sideband. At frequencies more than 1 kHz from the carrier, the noise floor of crystal oscillators is low enough to enable the generation of low-index frequency-modulation (FM) sidebands that are much higher than the noise floor. In this case they can be



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measured with a spectrum analyzer. Let the sideband level be $L(f_m)$. The spectral density of the modulated signal may be written as:

$$S_{\varphi m}(f) = S_{\varphi}(f) + 2L(f_m)\delta(f - f_m) \qquad (14)$$

$$\Delta t^2_{RMS_m} = \Delta t^2_{RMS} + \frac{T_0^2}{\pi^2} 2L(f_m)\sin^2(\pi f_m \tau) \qquad (15)$$

This equation makes it possible to calculate the modulation induced jitter versus time between measurements.

Equation 15 was checked using a 20.48-MHz high-performance complementary-metal-oxide-semiconductor (HCMOS)/transistor-transistor-logic (TTL)-compatible VCXO. The VCXO was modulated by a signal with a constant amplitude and frequency. The level of modulation was chosen to be low enough to excite only one FM sideband. The level of FM was controlled with an HP 8591E spectrum analyzer. The RMS jitter versus τ was measured by a Tektronix 11801A DSO equipped with an SD-24 sampling head. The measurements were conducted as follows:

The horizontal position of the DSO was set for time τ after the front of the first period (the trigger event). The RMS jitter was measured with and without modulation (Δt_1 and Δt_2). The measured modulation-induced jitter was then calculated as:

$$\Delta t_m^{meas} = \sqrt{\Delta t_2^2 - \Delta t_1^2} \qquad (16)$$

This procedure makes it possible to remove the DSO's contribution and also the unmodulated jitter of the oscillator.

The modulation-induced jitter was measured at a 10-kHz frequency with one level of modulation (L = –43 dB) and at a 20-kHz frequency with two levels of modulation (L = –43 dB and L = –40 dB). According to Eq. 15, the induced jitter of the 20.48-MHz VCXO modulated at 10 kHz with L = –43 dB is:

$$\Delta t_m^{calc} = \frac{1}{20.48 \times 10^6 \pi} \times \sqrt{2 \times 10^{-4.3}}\left|\sin\left(\pi 10^4 \tau\right)\right| \times 10^{12}\ ps \qquad (17)$$

This function reaches its maximum value of 155.6 ps at τ = 50 μs.

The plots of measured and calculated modulation-induced jitter are presented in Figs. 1 and 2.

The result of these measurements confirms the theory and demonstrates that Eq. 13 is valid.

Here we can point to a practical benefit of Eqs. 13 and 15. The phase-noise spectrum of oscillators, especially PLL-based oscillators, often contains strong deterministic (i.e., patterned or repeatable) spectral lines. These lines can be caused by

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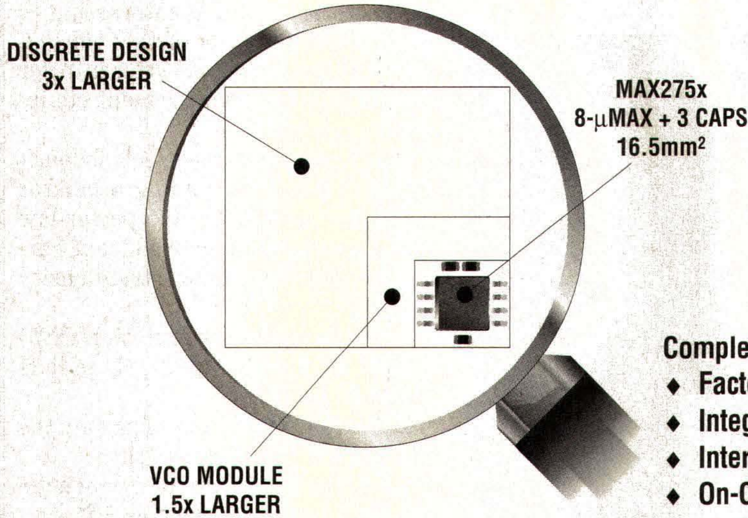

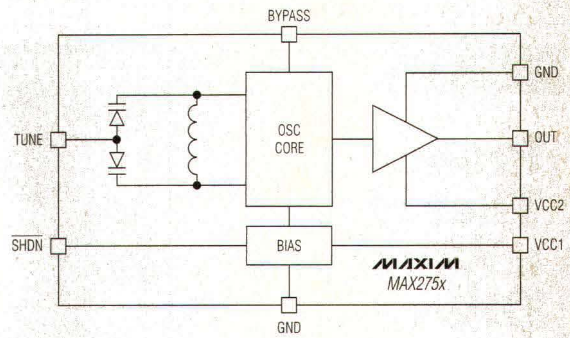

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power-supply noise, the PLL reference frequency, and other unwanted frequency-synthesis products. Equation 15 makes it possible to calculate the contribution of such spectral lines to the measured jitter for a given τ. By comparing the results for each spectral component to the measured oscillator jitter, it is possible to identify whether the jitter is induced by random phase noise or by certain deterministic spectral components.

CALCULATING JITTER

Now that the validity of the basic approach has been demonstrated, we will return to the question of calculating period jitter from phase noise. As mentioned previously, it is almost impossible to make a direct comparison of jitter calculated from measured phase noise and the jitter result provided by a DSO or any similar instrument due to the presence of amplitude noise and measurement-equipment time-base noise. The absence of data for effective bandwidth of the jitter-measurement equipment and oscillator wideband phase noise are additional complicating factors. Still, it is interesting to compare the results of this calculation with the measured jitter in order to estimate the measurement equipment's contribution.

It is known that the spectral density $S_\phi(f)$ of a free-running oscillator could be modeled by five power-law noise processes that produce a particular slope on the spectral-density plot:[3]

1. White phase modulation (PM) [white of phase]: the $S_\phi(f)$ plot is reported as$1/f^0$.
2. Flicker PM (flicker of phase): the $S_\phi(f)$ plot is reported as $!/f^1$.
3. White FM (white of frequency): the $S_\phi(f)$ plot is reported as $1/f^2$.
4. Flicker FM (flicker of frequency): the $S_\phi(f)$ plot is reported as $1/f_3$.
5. Random walk FM (random walk of frequency): the $S_\phi(f)$ plot is reported as $1/f_4$.

Estimating the integral in Eq. 13 in the presence of each of these processes:

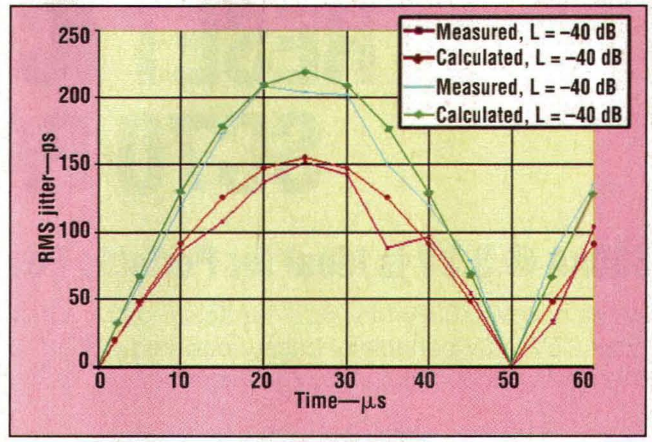
White PM:

$$\Delta t^2_{RMS_1} = \frac{T_0^2}{\pi^2}\int_0^\infty S_\phi(f)\sin^2(\pi f\tau)\,df = \frac{T_0^2}{\pi^2} S_{\phi WPM}\int_0^{f_h}\sin^2(\pi f\tau)\,df \quad (18)$$

Calculating the integral and substituting $\tau = NT_0$:

$$\Delta t^2_{RMS1} = \frac{T_0^2}{\pi^2} S_{\phi WPM} \times \frac{f_h}{2}\left(1 - \frac{\sin 2\pi N\frac{f_h}{f_0}}{2\pi N\frac{f_h}{f_0}}\right) \quad (19)$$

If $2\pi N = f_h/f_0 \geq 1$, the second term



2. Doubling the modulation levels and frequency from Fig. 1 results in the plots shown here. The two modulation levels are –43 and –40 dB, while the frequency is 20 kHz.

in the brackets in Eq. 19 is small compared to 1. It immediately follows that:

$$\Delta t^2_{RMS_1} \cong \frac{T_0^2}{\pi^2} S_{\phi WPM} \times \frac{f_h}{2} \quad (20)$$

In that case, the white phase-noise-induced jitter is not dependent on N and does not accumulate. This is because the white noise is uncorrelated and the second bracketed term in Eq. 9 equals zero.

If $2\pi N f_h/f_0 << 1$, Eq. 19 could be reduced to:

$$\Delta t^2_{RMS_1} \cong T_0^4 S_{\phi WPM} \times \frac{f_h^3}{3} \times N^2 \quad (21)$$

In that case the white phase-noise-induced jitter is proportional to the number of periods and does accumulate.

It also should be noted that if $f_h = f_0$, Eq 19 equals Eq. 20 for any N.

We see that f_h determines the jitter behavior in this case.

Flicker PM:

$$\Delta t^2_{RMS_2} = \frac{T_0^2}{\pi^2} S_{\phi FPM} \int_0^{f_h} \frac{\sin^2(\pi f \tau)}{f} df =$$

$$\frac{T_0^2}{\pi^2} S_{\phi FPM} \int_0^{\pi f_h \tau} \frac{\sin^2(x)}{x} dx \quad (22)$$

Evaluation of Eq. 22 shows that it is a slow-growing function of τ and f_h.

White FM:

$$\Delta t^2_{RMS_3} = \frac{T_0^2}{\pi^2} S_{\phi WFM} \int_0^{f_h} \frac{\sin^2(\pi f \tau)}{f^2} df =$$

$$\frac{T_0^2}{\pi^2} S_{\phi WFM} \times \pi\tau \int_0^{\pi f_h \tau} \frac{\sin^2(x)}{x^2} dx \quad (23)$$

(continued on page 157)

Design A Very-Wide-Range VCO

This voltage-controlled oscillator offers fast tuning characteristics, minimal hysteresis, and high modulation sensitivity while maintaining a small footprint.

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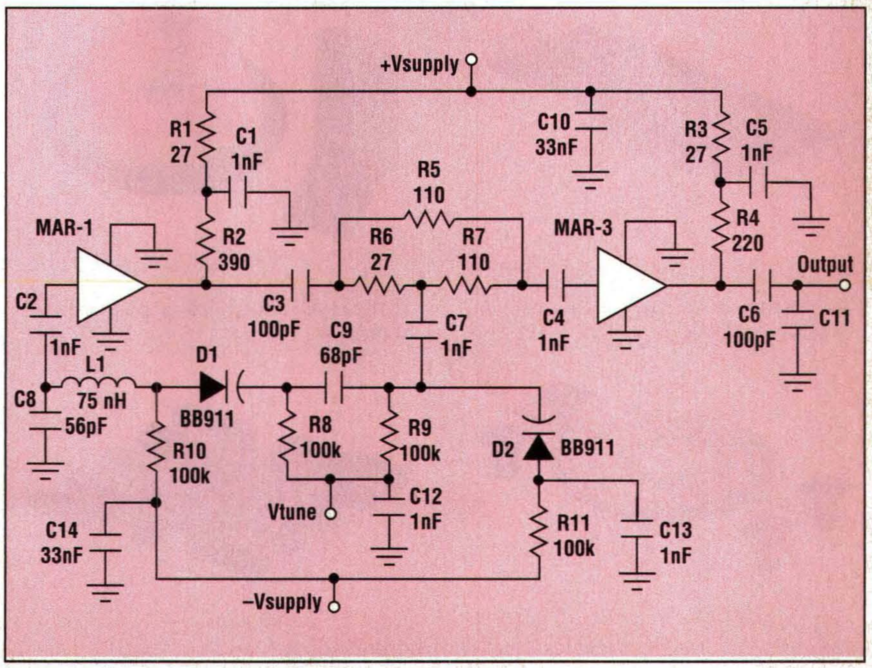
VOLTAGE-CONTROLLED oscillators (VCOs) are ubiquitous and pervasive components in the design of today's wireless systems. A very-wide-range voltage-controlled oscillator that exhibits good frequency and amplitude stability over its entire frequency span is an important component in a large number of RF and microwave applications. Wide-pull, phase-locked loops (PLLs), for example, incorporate a single, continuously tuned VCO and broadband sweep generators with good stability.[1,2]

But it is often difficult to meet the required specifications due to effects like frequency instability versus temperature, harmonics, and phase noise. The phase noise of the VCO is determined primarily by the overall Q of the circuit, the noise of the power-supply, and the external tuning-voltage supply. To design a circuit with high Q, the tuning bandwidth must invariably suffer.[3] Therefore, to achieve the overall per-



1. This schematic of the very-wide-range VCO shows the two monolithic amplifiers, MAR1 and MAR3.

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IE3D Simulation Examples and Display

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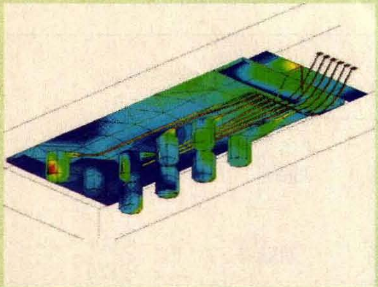
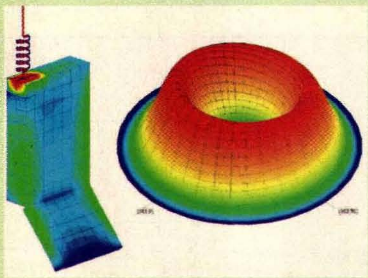
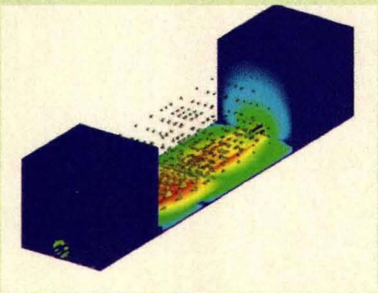


The current distribution and radiation pattern of a handset antenna modeled on IE3D

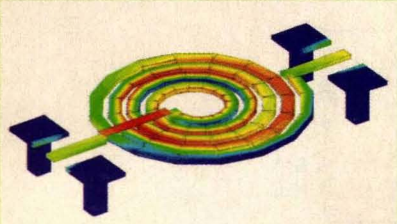


FIDELITY Examples

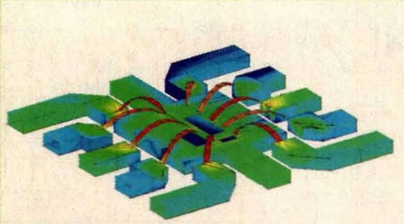
The near field and Poynting vector display on a packaged PCB structure with vias and connectors



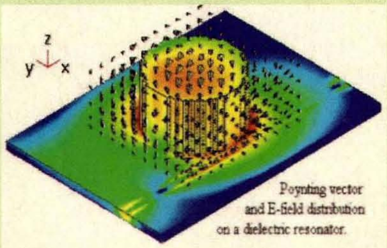
IE3D modeling of a circular spiral inductor with thick traces and vias



IE3D modeling of an IC Packaging with Leads and Wire Bonds



FIDELITY modeling of a cylindrical dielectric resonator and the Poynting vector display



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filter's S-parameters—that is, with S-Parameter Controller, "SP1" (Fig. 3). However, in order to cosimulate X2 within the platform (Fig. 1), the controller must be an envelope controller. Its timing parameters should be synchronized with the rest of the platform. That is, the "step" value of envelope controller is:

Step = 0.5 × Symbol_Time/Ratio

where:

Symbol_Time is the length of a symbol, and

Ratio is an up-sampling value.

The actual step value is reported as "tstep" in the Variable and Equations component "VAR1" in Fig. 1.

The remaining components of Fig. 1 are sinks and a cosimulation controller.

The sinks measure the spectrum at the input and the output of the filter, measurements of voltage (to measure input and output RF powers and output-constellation diagram) and measurements of EVM.

SIGNAL SPECTRUM

The signal spectrum in ADS is determined by applying a Fast Fourier transform (FFT) to the signal in the time domain. It is important to find an optimum value of FFT resolution (determined by the number of points) and the value of averaging. The number of points must be large enough to produce a correct representation of a signal in the frequency domain, and at the same time must not be too large in order to avoid the confusion of placing a data marker in an interpolated position while assuming a real value.

It was found that an FFT with a number of points = 1024 and a value of averaging = 25 gives consistently good results coinciding with those using a much-larger number of points and different values of averaging. The values of input and output average powers are calculated from the signal's spectrum content. The input spectrum is determined on the sink F_IN, and the output spectrum on F_OUT (Fig. 1).

The voltages (and consequently the instantaneous power levels) are determined as a result of straightforward measurements on the time sinks. The input instantaneous power is determined on the sink T_IN, the output instantaneous power—on T_OUT (Fig. 2). Sinks T_EVM_Q and T_EVM_I are used to determine an output-constellation diagram (Fig. 1).

The EVM is determined on its own sink, E1 (Fig. 1). To use it properly, a filter's output RF signal is demodulated on QAM demodulator Q1, and the I and Q bit stream from the output of Q1 is applied to root-raised-cosine (RRC) filters L1 and L2. The resulting signals are applied to E1. It is important to find correct timing values for E1, since the EVM calculation in ADS is performed by aligning the input and output bits. The sampling of the signal should be performed at the optimal symbol's timing phase. The delay of the library models of RRC filters (Figs. 1 and 2) was chosen to be 8*Symbol_Time μs. To compensate for that delay in the RRC filters in the signal source and the demodulator, the sampling of the output signal should begin in the middle of the first symbol delayed by 16*Symbol_Time μs. To adjust for the up-sampler (U1, Fig. 2), which is used to refine measurements and to increase the frequency span, the signal sampling should begin a little earlier, by the ratio of a quarter of a Symbol Time to the up-sampling value. If there are any other delays (such as a group delay of a filter) they should be accounted for by introduction of their own delay values. That is, the resulting "start" value of E1 is:

Start = 16.5*Symbol_Time—0.25*Symbol_Time/Ratio + Delay

where:

"Symbol_Time" is the length of a symbol,

"Ratio" is an up-sampling value, and

"Delay" is the value of all the additional delays.

The "stop" value of E1 is less critical. It should be large enough to conduct meaningful statistical calculations. The rest of the E1 parameters are self-explanatory.

The cosimulation controller O1 is required, since ADS has two different simulation engines for RF and for digital-signal-processing (DSP) calcula-

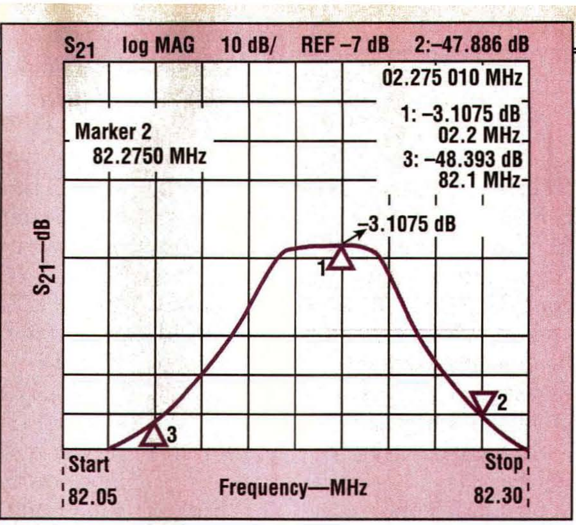
tions. The function of that controller is to provide a seamless transition between each of the engines. Its only parameter, output frequency, should coincide with the frequency value at the envelope controller of X2.

The π/4DQPSK modulated signal generated in X1 (Fig. 2) is applied to the filter's (X2), which is an S-parameter file obtained as a result of actual measurements (Fig. 3).

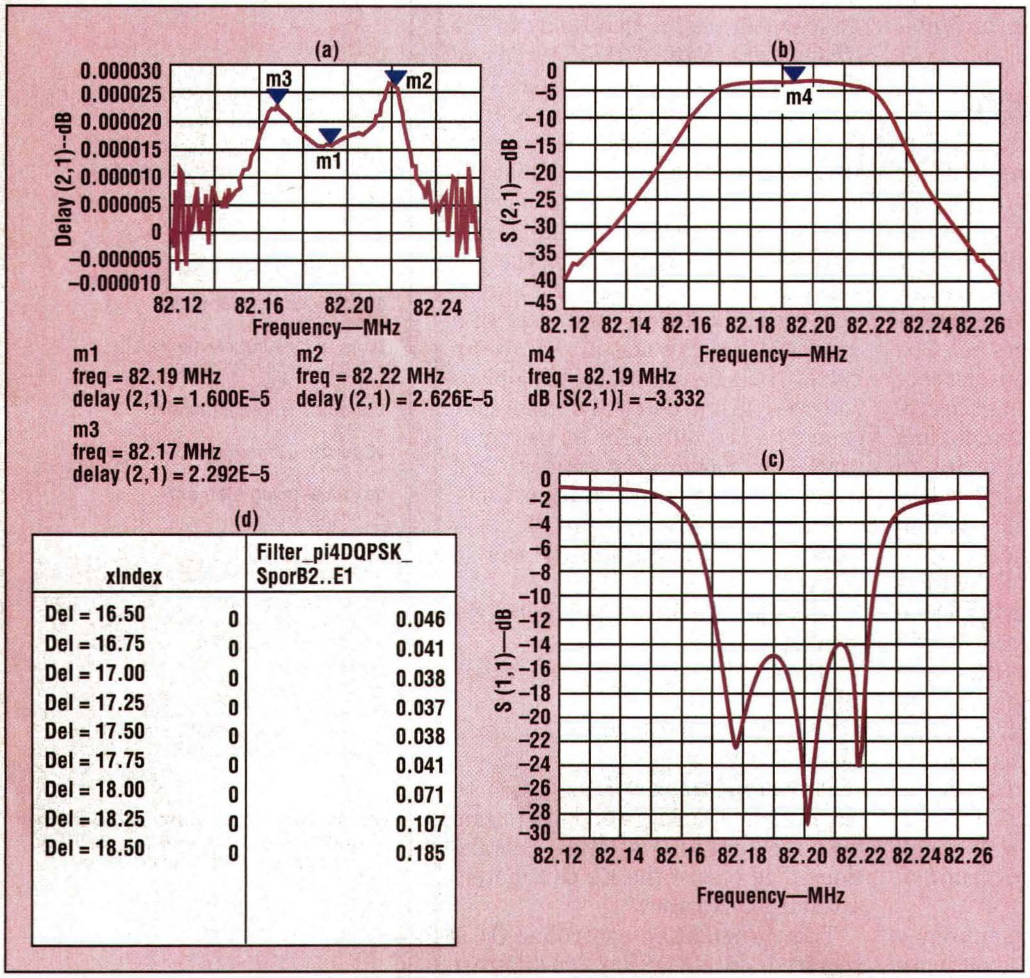
At first the S-parameters and group delay are plotted. To obtain the plots, filter configuration (Fig. 3) is simulated with the S-parameter controller. The parameters of this controller are straightforward, however the group-delay simulation has to be activated. The initial value of a group delay is obtained from the plot. However, that value is frequency dependent. It is necessary to determine some kind of an averaged group delay; that is, the resulting value of a group delay affecting all the frequency components of a Gaussian-shaped input signal. The process of precise determination of that value is cumbersome. There is a better way, however. It is based on the ADS method of EVM calculation.

As mentioned above, the EVM calculation in ADS is performed by aligning input and output bits; that is, by compensating all the delays caused by the system. Those delays that are impossible to compensate are those that cause EVM degradation. In order to find the required averaged group-delay value, the initial value is swept, the values of EVM are observed, and its minimum is found. The resulting value is the sought value of EVM that has been caused by the filter's group delay.

All simulated plots are illustrated in Fig. 4. As shown in Fig. 4d (table), the minimum value of EVM is 0.037 (3.7 percent).



5. This S-parameter curve illustrates the S_{21} measurement taken from an actual bandpass filter having the same characteristics as the simulated filter. Compare this curve with that of Fig. 4b to see the similarities.



xIndex		Filter_pi4DQPSK_ SporB2..E1
Del = 16.50	0	0.046
Del = 16.75	0	0.041
Del = 17.00	0	0.038
Del = 17.25	0	0.037
Del = 17.50	0	0.038
Del = 17.75	0	0.041
Del = 18.00	0	0.071
Del = 18.25	0	0.107
Del = 18.50	0	0.185

4. These plots [(a) through (c)] represent the results of the ADS simulation and the calculation of the EVM value (0.037) in the table (d). The group-delay time-response curve is shown in (a), and S-parameter curves are shown for S_{21} (b) and S_{11} (c).

THE CONFIRMATION

To confirm the validity of the simulation, the simulated data needs to be compared with experimental data. Experiments were conducted on a bandpass filter manufactured by Samsung (model No. H82AC) operating at a center frequency of 82.2 MHz. S-parameters and group delay were measured on an HP8753D network analyzer. EVM was measured on an Anritsu test setup: A TDMA-modulated signal from the output of an MS3670B generator was fed into the filter and its output EVM was measured by a

(continued on p. 157)

formance and to select proper VCO specifications for critical applications, one must evaluate these devices accurately.

This article presents the design of a VCO employing varactor-diode tuning for applications where fast tuning characteristics, minimal hysteresis problems, high modulation sensitivity, low power consumption, and small size are needed.

VCO SPECIFICATIONS

The VCO has been designed to meet the following specifications:

- V_{supply}: +12 VDC ±10 percent.
- I_{supply}: 55 mA maximum.
- Tuning voltage (V_{tune}): –10 VDC to +8 VDC.
- Frequency-tuning characteristic (frequency versus tuning voltage): 10 to 12 MHz/VDC.
- Output power (fundamental sinusoidal frequency output of the oscillator measured into a 50-Ω load): +8 dBm, minimum.
- Operating temperature range: –10 to +60°C.
- Frequency versus temperature (variation of frequency with temperature at a fixed tuning voltage): 57 kHz/°C, maximum.
- Harmonics: ±20 dBc.
- Size: 0.8 × 0.8 × 0.4 in. (2.03 × 2.03 × 1.02 cm).

THE VCO CONTAINS TWO MONOLITHIC AMPLIFIERS, MAR-1 AND MAR-3, WHICH PROVIDE VERY FLAT, WIDE-BAND RESPONSE. MAR-1 IS USED AS THE OSCILLATOR, WORKING IN A COLPITTS CONFIGURATION. MAR-3 IS AN ADDITIONAL AMPLIFIER THAT BRINGS THE OUTPUT POWER UP TO THE REQUIRED LEVEL.

Figure 1 shows the schematic of the VCO, together with some component values and data on the amplifiers used in the circuit. The VCO contains two monolithic amplifiers, MAR-1 and MAR-3, which provide very flat, wide-band response. These amplifiers are used because they have wide bandwidth and high output gain. MAR-1 covers the frequency range from DC to 1 GHz. Between 100 and 1000 MHz, its typical gain ranges from 18.5 to 15.5 dB. MAR-3 covers the frequency range from DC to 2 GHz. Between 100 MHz and 1 GHz, its typical gain ranges from 12.5 to 12.0 dB.

MAR-1 is used as the oscillator, working in a Colpitts configuration. MAR-3 is an additional amplifier that brings the output power up to the required level. Resistors R_1 and R_2 and capacitors C_1, C_2, and C_3 bias

MAR-1. R_3, R_4, C_4, C_5, and C_6 bias MAR-3.

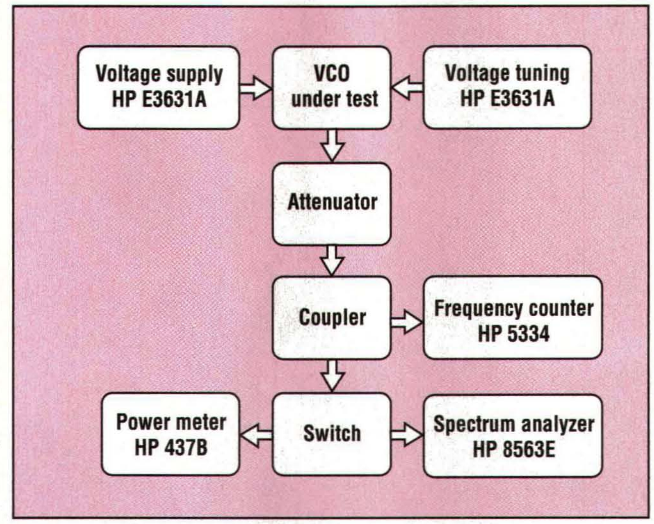
TUNING DIODES

The selection of the proper tuning diodes is important as the proper way of connecting them. In this circuit two BB911 very-high-frequency (VHF) variable capacitance diodes have been employed. These diodes are characterized by their high linearity in the tuning range.

When building the prototype, one should take the following considerations into account to obtain the best overall performance from the previous specifications.

The VCO must be set on a ground plane with a number of plated through holes (vias) spaced at intervals of between 20 and 30 mils. The power-supply and tuning voltage must be connected to the printed-circuit-board (PCB) ground plane and all VCO ground pins must be soldered directly to the printed ground plane. It is necessary to avoid the presence of soldermask, which can sometimes degrade VCO performance. Furthermore, for good RF grounding, several decoupling capacitors must be used.

Figure 2 shows the VCO test setup. In this configuration, the VCO must be located as close as possible to the attenuator, thus avoiding any adverse effects due to load mismatch. The supply voltage and tuning voltage present low noise to avoid fluctuations in frequency and power. ●●



2. This block diagram shows the components used in the VCO test setup.

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Slotted-Line System Measures S-Parameters Automatically

A microcomputer-controlled slotted line provides a rapid and accurate method for measuring scattering coefficients.

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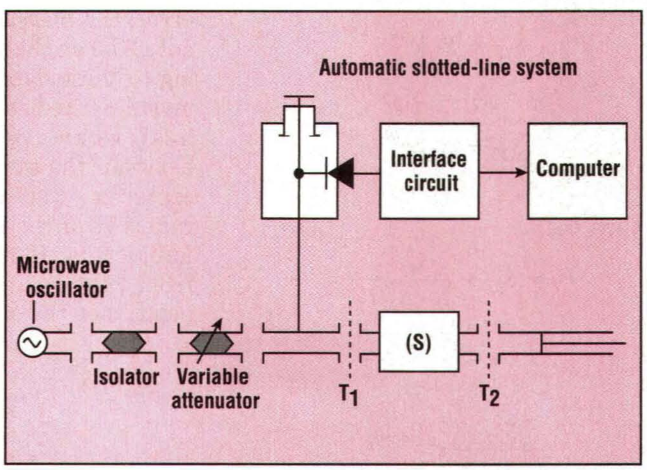
MANY microwave components and elements are two-port networks—joints, bends, irises, matching screws—whose characteristics can be expressed by scattering coefficients (S parameters). In general, two measurement methods are used to measure these coefficients: the three-points method and the shorting-plunger method, with the latter more commonly used.

Using the conventional shorting-plunger method to measure the S parameters of a reciprocal, nonloss two-port network, the measured data must be plotted individually on a Smith chart to produce a smooth circle of the input-reflection coefficient Γ_1. The center C and radius R of the Γ_1 circle, the center 0 of the Γ_2 circle, the image-circle center O, and the arguments θ_{12}, θ_{22} of the S_{12}, S_{22} parameters, respectively, must be determined. Finally, the scattering parameters can be evaluated according to the following formulas:

$$S_{11} = \overrightarrow{OO'},$$

$$S_{22} = \frac{\left|\overrightarrow{O'C}\right|}{R} e^{j\theta_{22}},$$

$$S_{12} = \sqrt{R\left(1 - |S_{22}|^2\right) e^{j\theta_{12}}} \qquad (1)$$



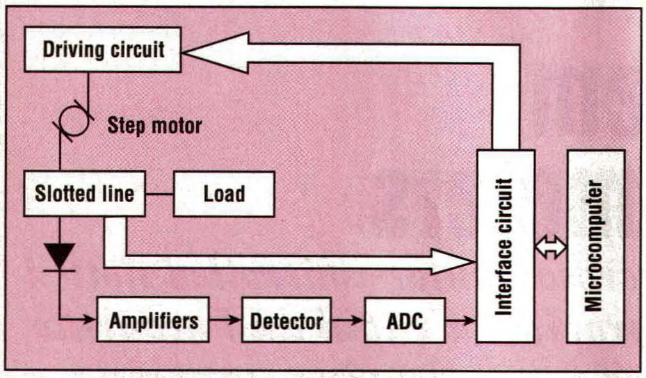
1. This is the shorting-plunger measurement setup with [S] representing the network whose S coefficients are to be measured.

Unfortunately, this measuring and graphing process is lengthy and tedious. Moreover, the statistical and mean method cannot be used, and the measurement accuracy is low.

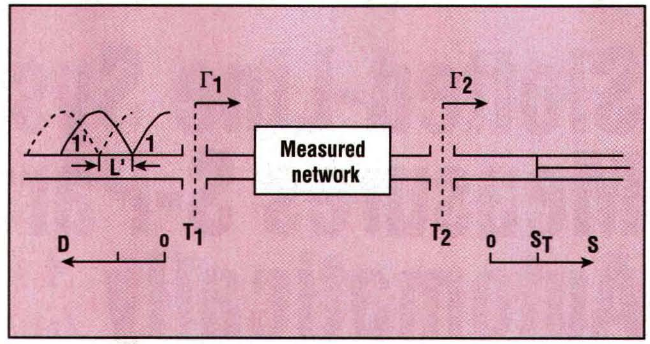
A better shorting-plunger method can be realized by an automatic slotted-line controlled by a microcomputer programmed with Turbo-C language and using the numerical-value method. This method offers advantages over conventional methods such as a simple setup, as well as rapid and accurate measurements.

THE SHORTING PLUNGER

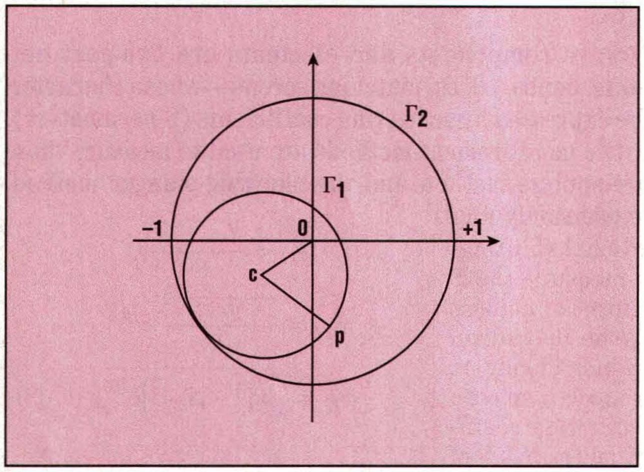
The shorting-plunger/numerical-value method is based on the graphic method and is abstracted into a mathematical mode. The measured data is processed by the microcomputer to find the optimum circle of the input-reflection coefficient and the image center about the Γ_2 circle, and to evaluate the S parameters under the statistical condition. In the measuring setup (Fig. 1), [S] repre-



2. The automatic slotted-line system enables the different positions of the shorting plunger in the slotted line to be acquired and measured under the control of the microcomputer.



4. Phase measurements are performed by moving the shorting plunger either toward or away from the measured network. This movement changes the wavelength as well as the phase angle.



3. The Γ_2 circle shown here enables another circle, Γ_1, with a different center and radius, to be generated. These coordinates are used to calculate the measured network's S parameters.

sents the measured two-port network; Γ_1 and Γ_2 are the input- and output-reflection coefficients of the measured network, respectively, and D_T and S_T represent the place readings of the reference planes corresponding to the relative ports, respectively.

The block diagram of the automatic slotted-line system is shown in Fig. 2. The motor is controlled by the microcomputer and drives the probe to move back and forth along the slotted-line. The detected signal is fed into the microcomputer interface circuit through the amplifiers and the analog-to-digital converter (ADC). With this automatic slotted-line system, the following parameters can be automatically acquired and measured: the wavelength guide λ_g, the place reading, D_T of the input-reference plane of the measured network, the N-node places reading D_{mini} of the standing wave, and the N VSWRs S_i corresponding to the N different-places readings L_i of the shorting plunger. Thus, provided the application software is designed to process the data in accordance with the method of numerical value, the Smith chart can be drawn and the scattering parameters can be evaluated.

Now, suppose the S parameters are written as:

$$S_{11} = |S_{11}|e^{j\theta_{11}},$$
$$S_{12} = S_{21} = |S_{12}|e^{j\theta_{12}},$$
$$S_{22} = |S_{22}|e^{j\theta_{22}} \quad (2)$$

then[1]

$$\Gamma_1 = S_{11} + \frac{S_{12}^2\Gamma_2}{1 - S_{22}\Gamma_2} = S_{11} + \frac{S_{12}^2 S_{22}^*}{1 - |S_{22}|^2} + \frac{S_{12}^2}{1 - |S_{22}|^2} e^{j(2\phi - \theta_{22})} \quad (3)$$

where:

$-\Phi$ = the argument of $(1 - |S_{22}|e^{j(\Phi + \theta_{22})})$, while $\Phi = \pi - 2\beta(L - L_T)$, and $L - L_T$ = the distance moved by the plunger.

It is thus clear that Eq. 3 traverses a circle or a set of circles on the complex plane Γ_2 and another circle or another set of circles on the complex plane Γ_1. Hence, it is possible to select N values of Γ_2, measure N corresponding to Γ_2 values which are clearly arranged on a circle on the Γ_1 complex plane. However, selecting N values of Γ_2 is actually realized by moving the shorting plunger N times (N is an even number) within $\lambda_g / 2$. So when the plunger is moved continually, Φ changes continually, and 2ϕ changes follow them. Then the circle of Γ_1 is as shown in Fig. 3. Using the radius and the center coordinates of the Γ_1 circle, the S parameters can be calculated.

However, due to measurement error, the measured data of Γ_{1i} cannot all be on the same circle. According to these data, an optimum circle must be fitted so that the $\Sigma\delta_i^2$ is minimal, where: δ_i is the difference between the distance from circular center $(x_c\ y_c)$ to every point and the radius R of the fitted circle. Therefore, letting R_i express the distance from the circular center to every point, then the object function is:

$$G(R, R_i) = \sum_{i=1}^{N} (R - R_i)^2 = \sum_{i=1}^{N} \delta_i^2 \quad (4)$$

To evaluate the extreme value for

$G(R, R_i)$, let $\partial G/\partial R = 0$. The following relations can be derived:

$$R = \frac{1}{N}\sum_{i=1}^{N} R_i \qquad (5)$$

$$G = \sum_{i=1}^{N} R_i^2 - \frac{1}{N}\left(\sum_{i=1}^{N} R_i\right)^2 \qquad (6)$$

By designing the application program with the microcomputer and using the technology about the optimization, the circular center coordinates (x_c, y_c), and radius R can be calculated, while the object function $G(R, R_i)$ is minimized.

Furthermore, according to the measured data of Γ_1, using the numerical-value method and straight-fitting method, the coordinates (x_o', y_o') of the image-circle center O′ (i.e., the image of the Γ_2) circle center) can be found. Then, it can be determined that:

$$S_{11} = \overrightarrow{OO'},$$

$$|S_{22}| = \frac{\overrightarrow{|O'C|}}{R},$$

$$|S_{12}| = \sqrt{R\left(1 - |S_{22}|^2\right)} \qquad (7)$$

As for the arguments of S_{12} and S_{22}, they will be submitted to the following relations:

$$\theta_{12i} = tg^{-1}\left(\frac{y_i - y_{o'}}{x_i - x_{o'}}\right) - \frac{1}{2}tg^{-1}\left(\frac{y_i - y_c}{x_i - x_c}\right) - \frac{1}{2}\phi_i + \alpha \qquad (8)$$

where:

$x_i - x_o' < 0$ and < 0, then $\alpha = \pi/2$, or $x_i - x_o' > 0$ or $x_i - x_c > 0$, then $\alpha = 0$.

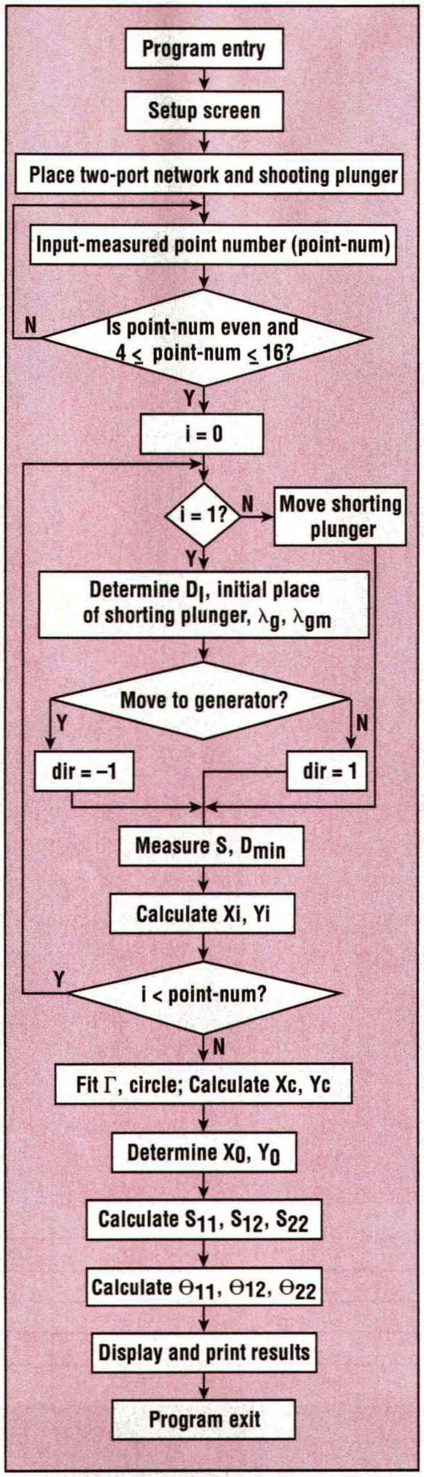
$$\theta_{12} = \frac{1}{N}\sum_{i=1}^{N} \theta_{12i} \qquad (9)$$

$$\theta_{22} = 2\theta_{12} - tg^{-1}\frac{y_c - y_{o'}}{x_c - x_{o'}} \qquad (10)$$

Hence, calculate every θ_{12i}, and then evaluate their mean values using the microcomputer, in accordance with Eq. 8. Then θ_{12} and θ_{22} can be calculated.

At this point, all of the S parameters have been evaluated.

The process described can be realized by an automatic slotted-line and the numerical-value method. The test, data acquisition (DAQ), data processing, calculations, plotting, and results can be displayed and printed at one time by running the application software. Thus, the goal of a fast and convenient measurement scheme is achieved.



5. This flow chart of the slotted-line measurement system begins with the shorting plunger in an initial position; as it moves to different positions, different phase angles Φ are produced.

THE FLOW CHART

In this measurement, the initial position of the shorting plunger can be determined as follows (Fig. 4): place the plunger in a starting position and determine the standing-wave node 1 that is the nearest node to T_1 by the automatic slotted-line method. This point is the reference plane about the plunger's initial position. After this time, whenever the shorting plunger is moved, the node also moves in the following manner: measure the distance L′ between the first node 1′ and point 1. Then the distance moved by the plunger is L = {λ g}/2 − L′ (if the plunger moves toward the terminal), or L = L′ (if the plunger moves toward the oscillator). The corresponding phase is $\Phi = \pi - 2\beta L$. Therefore, it is not necessary to measure the reference plane of T_2.

In the program flow chart (Fig. 5): D_T represents the position of the input-reference plane of the measured network; S and D_{min} represent the VSWR and the position of the node, respectively, which are used to find the corresponding reflection coefficient Γ_1; λ_g and λ_{gm} express the wavelength guides of the slotted and nonslotted waveguide, respectively. dir indicates the direction of movement of the shorting plunger. The distance L of the plunger and the phase, Φ, can be calculated in the program according to the value of dir. The method for evaluating x and y is the same as that used to evaluate the impedance.[2] The displayed and printed results include the output Smith chart and all of the measured results of the S parameters.

The application software consisting of motor driving, DAQ, data processing and calculation, chart drawing, and outputting and printing the results is designed with the Turbo-C language. It is simpler than using assembly language to design the

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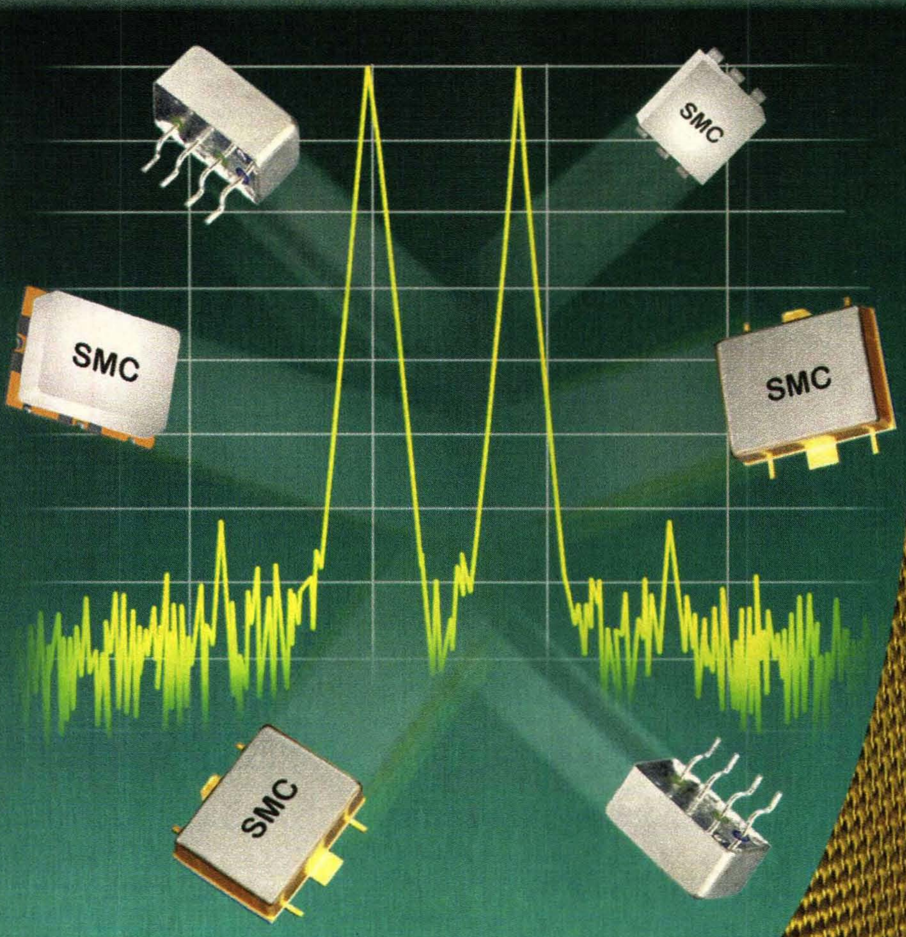

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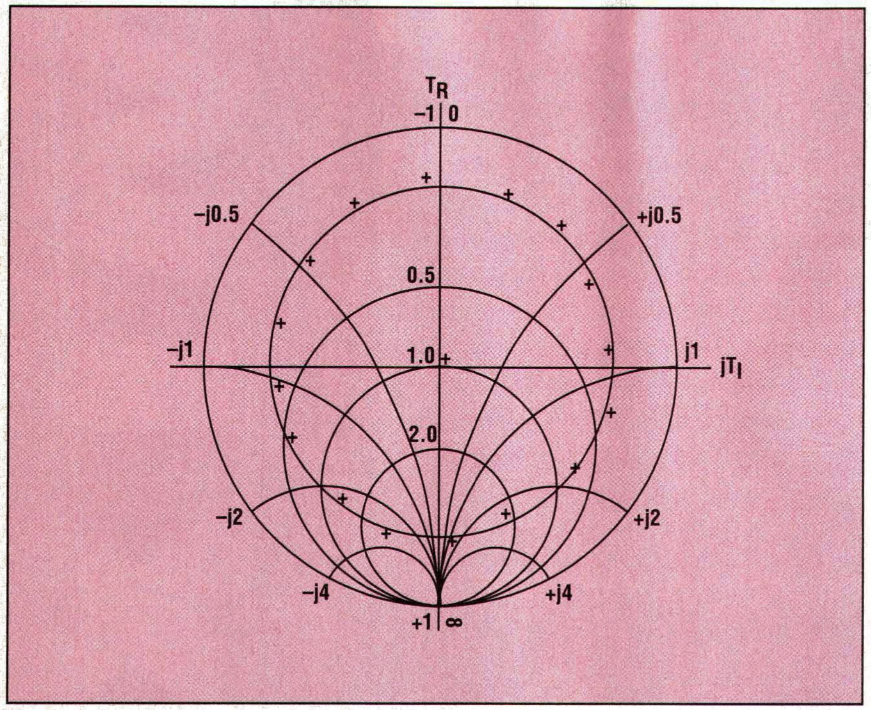

6. A Smith chart such as the one shown here displays the S parameters of the shorting-plunger measurement system. In this case, the results reflect 16 measured points.

motor drive and DAQ programs.

The S parameters of a variable attenuator as a two-port network were measured under the condition that the amount of attenuation is less, and the measured points are 4, 8, and 16, respectively. The three results show no basic difference. Thus, only the result in which the measured points are 16 is shown in Fig. 6. ••

Acknowledgement

The authors would like to thank Pan Chuhua, senior engineer of our department for providing some of the instrumentation for our experiments.

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For further reading

Pan Chuhua, *Microwave Experimental Teaching Materials,* Dept. of Electronics of Zhongshan University, August 1994 (China).

Selecting A Shielding Supplier

Part 1 of 2 parts

Keeping the supply line open to high-quality shielding materials is not simply a matter of picking the largest or lowest-cost vendor.

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SHIELDING materials gain in importance with the continued growth of wireless communications. As RF-based electronic devices proliferate, they must be designed and manufactured to co-exist without causing interference. Proper shielding can ensure such electromagnetic compatibility (EMC). Selecting a capable supplier of shielding products can also contribute to the optimum integration of shielding products into a final design.

Not all shielding suppliers are the same or offer the same level of services, expertise, products, or attention to detail. Dealing with the wrong supplier can be financially costly due to late deliveries and inferior product performance. What follows is a list of questions that can guide an engineer or procurement technician through the process of evaluating a supplier of shielding products.

Is the size of the shielding supplier important? Smaller companies can generally provide personal "one-on-one" customer service. That means—whatever questions or problems an original-equipment manufacturer (OEM) may have regarding the shielding products and whenever that help might be needed—the person responding to an inquiry answers it promptly and accurately, rather than deferring the question to other personnel within the company. Larger companies, due to their extra layers of departments, personnel, and large customer base, often cannot provide timely answers to inquiries, especially for those critical rush requests. Smaller customers that lack a history with a shielding supplier, or a large-enough order, may find themselves waiting at the end of a long line. A good rule of thumb when selecting the "right-sized" shielding company is to target a firm with between 50 and 150 employees and approximately $10 to $25 million in annual sales. Such companies should be small enough to provide personal service when needed, but large enough to provide the in-depth professional expertise, versatility of product line, and attention to needed for even those OEMs with diverse electronic-product lines.

1. A shielding supplier with a diversified product line can satisfy a greater range of customer requirements than a supplier with a limited product line.

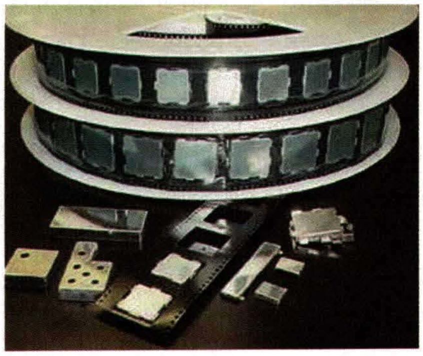
Is the length of time that a shield-

ing supplier has been in business important? The length of time that a company has been in business not only indicates its longevity, but also indicates its reliability. Companies that have been in business for a decade or more survive due to their ability to consistently satisfy their customers. However, many startup companies or industry newcomers, albeit important to the economy, can disappear within one to two years due to such factors as inadequate capitalization. It takes time for a company to build credibility. The time to experiment with a new supplier is not when a critical project is due.

What kind of products does the shielding supplier manufacture? Does a shielding supplier have a diversified or limited product line? Do they have the right product line to satisfy a wide range of application problems, or do they sell a limited off-the-shelf product line offering no opportunity for modification? There may be times when conventional shielding materials will not solve a unique application problem. Having a supplier with a diverse product line and the ability to modify their products without excessive artwork rework, tooling, or time delays could be critical to completing an OEM's project (Fig. 1).

Does the shielding supplier have a design program in place? Generally, every shielding company has technicians or designers on staff. However, the major difference between companies is that they may or may not have a true design program. A truly organized "program" will spell out who, what, when, where, how, and why on every inquiry for every designer. How are customers' inquiries supposed to be handled? Are their technicians assigned to a list of specific customers enabling the designers to become familiar with their customers' procedures, shipping requirements, product lines, specifications, etc? When is a response to an inquiry considered timely or untimely? Does someone evaluate the employees continuously? In a legitimate "program" the process of handling inquiries is methodically planned and its people are continuously trained. A customer's chance of getting a timely and intelligent answer to an inquiry is enhanced when working within a true design "program."

What are the qualifications of the shielding company's design staff? Qualifications are sometimes misleading since a college degree (in any discipline) does not always translate into a practical and timely answer to an inquiry. It is certainly helpful and an added value for any shielding company to have an electrical and mechanical engineer on staff as back up for those difficult, intricate, and unique design problems needing that extra boost of intelligence. But, do not underestimate the value of the technician with decades of practical experience in all facets of shielding design. Weigh the merits of each technician. Do they understand a problem? Are their solutions practical and timely? Are their specifications and drawings comprehensive enough to include all needed details and yet simple enough to follow, and practical to manufacture?



2. If a customer's requirement calls for a large quantity, a shielding supplier should be able to offer tape-and-reel capabilities.

What type of computer-aided-design (CAD) capabilities does the shielding supplier have? Every supplier has CAD capabilities, but are they compatible with a customer's requirements? Does a supplier use the same software as the customer, and will the supplier be able to work with the customer's files?

What is the availability of the shielding supplier's resource materials? What types of supplies are generally needed, and what is kept on hand by a particular shielding supplier? What is their general inventory of those stocks? Are any particular materials, such as beryllium copper (BeCu), in short supply? How fast can a supplier obtain a sufficient amount of stock for an order? It is obviously advantageous working with a supplier who has sufficient stock or knows how and where to get it—resourcefulness is a big plus when the stock for an order is in short supply and one's productivity could be seriously slowed or halted.

What kind of customer-service program does a shielding supplier have? This is one of those key elements in evaluating suppliers properly. How well are customers treated when they call? What is learned on the first call? Are samples offered on the first call, or has an issue become confused due to misunderstandings on the part of the shielding supplier's customer-service people? Is it necessary to call back several times in order to get a person on the line rather than voice mail? And is there a long wait for a return call? A good customer-service "program" instills consistency, accuracy, and diplomacy.

Since the timing in getting a finished product to market is often critical, and because shielding is often an afterthought, it is important to reach an actual person on the first call. A conscientious supplier serving this time-sensitive industry should use voice mail sparingly. If a customer's normal contact at a shielding supplier is not available, someone else should assist the customer with the same concern and equal dexterity. When a customer-service program works well, a design solution can usually be agreed upon and samples shipped in the same day. When it does not work, it could take days to reach a solution, and weeks to receive the samples.

How long does it take for a shielding supplier to respond to an inquiry? Besides the design-solution response time, there is also the critical timing for processing information that could be a response to advertising and direct mail. Offerings such as catalogs, sell sheets, tutorial articles, sample packages, etc. should be sent

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expediently. First Class postage should be the norm rather than the use of bulk-rate postage. Some companies can take up to six weeks to deliver literature.

One important factor in supplier response, especially with the largest shielding suppliers, is the "small-fish-in-the-big-pond" syndrome. At a larger shielding supplier, a customer in need of a 30-piece order will not receive the same treatment as a customer with a 300,000-piece order. Smaller quantities usually imply the need for a smaller shielding supplier in order to receive proper customer treatment. Better still, since needs can change instantly from low to high production runs, the ideal supplier is one that can handle the full spectrum of customer requests, but with equal enthusiasm for small as well as large quantities (Fig. 2).

Does a shielding supplier deliver on time? Another very important key in evaluating a supplier. What percentage of their deliveries arrives on time? Do they monitor or have a tracking system for deliveries? Delivering a custom product late can have a serious effect not only on delivery of the finished product to one's customer, it could cost one's company an account. Do they deliver one's samples, prototypes, and production orders consistently on time? Some larger companies can take four-to-six weeks to deliver samples.

Is a shielding supplier willing to provide customer references? An OEM should ask for references or testimonials from any new supplier. If they have none or only a few, one should seriously doubt their ability to satisfy their customers. Ask for at least five-to-10 references. Obtain recent references within the current or prior year. If a company has operated for 10 years and does not have five-to-10 satisfied customers, stay away from them.

How good is the quality of a shielding supplier's samples? Samples are a window into the supplier's ability to produce a quality product. If the quality of their samples is poor, then their product will probably be manufactured with the same poor quality. Samples should be received in good condition and should reflect the solution recommended by the technician at the shielding company.

The second part of this article will continue with the question-and-answer format presented here. The questions to be considered advance the theme of selecting a shielding supplier, with an emphasis on some of the business aspects of the selection process. For example, the article will offer answers to the questions of how good a marketing communications operation that the supplier should operate (literature, website, etc.) and what kind of resources the supplier provides for resolving disputes should there be questions regarding delivery, price, and other business matters. It will also tackle the subject of contracts, including invoicing, pricing, and credit arrangements. Finally, Part 2 will deal with the question of product quality. ••

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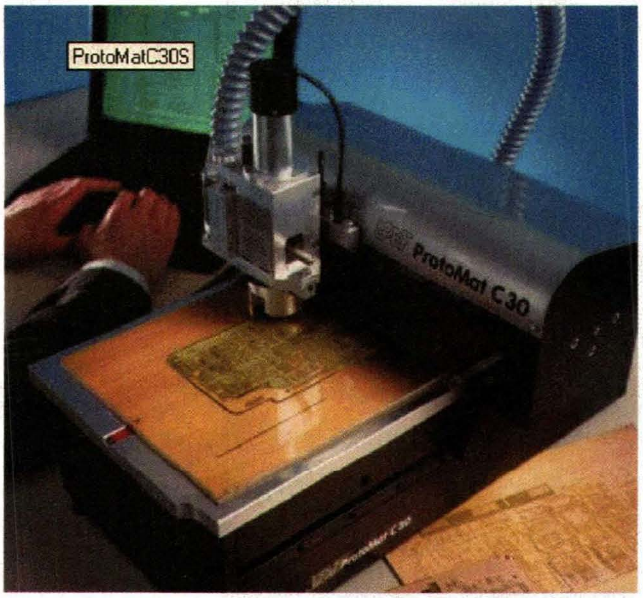
ADVANCES in desktop circuit-board routers have extended the speed, safety, and convenience of mechanical printed-circuit-board (PCB) prototyping to include the most demanding applications. Unique pneumatic systems control the cutting process more precisely and gently than previous methods had, making it possible to create circuits at the desktop on highly sensitive Teflon substrates such as RT/duroid®.

Modern circuit-board routers such as the portable machines available from LPKF (Wilsonville, OR) are capable of producing structures with track widths as fine as 100 μm, with precise cutting channels. These machines can achieve accuracy of better than 0.2 mil to ensure the faithful reproduction of fine-pitch structures and high-density circuits. One of the key advances in these machines is the use of adjustable-speed three-phase motors capable of operating to 100,000 rpm. By employing such high spindle speeds, greater geometric precision is possible, while also extending the life of the cutting tools.

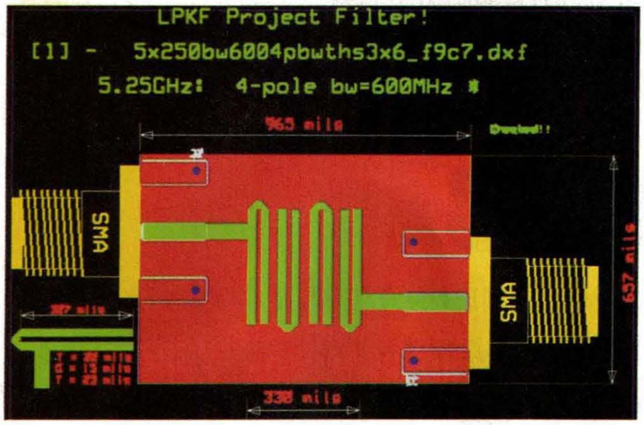


1. Compact table-top circuit-board routers can provide high precision, repeatability, and fast turnaround times on circuits fabricated on Cu-clad Teflon substrates.

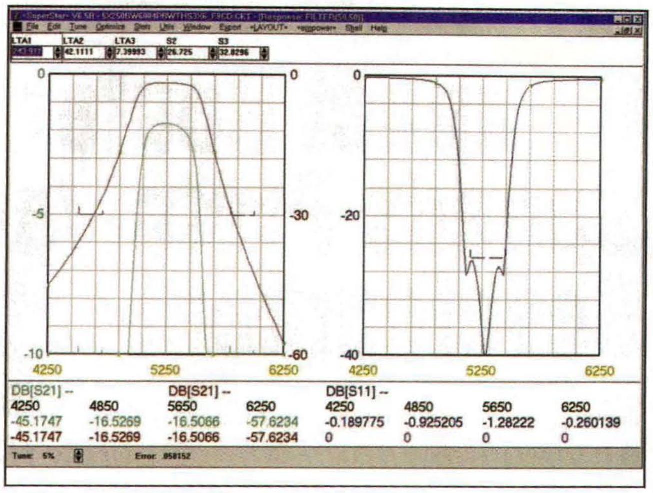
These compact routers (Fig. 1) can be used anywhere in a high-frequency electronics laboratory or production facility. Once designed solely for single-layer designs, they can now fabricate multilayer circuits with the aid of accessory presses and through-hole systems. Typical four- and six-layer circuit boards can be produced in a few hours.

GETTING THE EDGE

Since time to market and design integrity are critical factors in RF wireless communications markets, the availability of these tabletop circuit-board routing systems can provide a competitive edge for companies seeking to supply fast turnaround times and custom solutions. Producing one's own prototype circuits (rather than subcontracting the work) through mechanical milling also helps protect intellectual



2. A four-pole bandpass filter for wireless communications in the 5.2-GHz band was selected for the purposes of comparing mechanical milling and chemical-etching fabrication processes.



3. This analysis of filter behavior was performed with the help of the M/FILTER software.

property (IP) without resorting to potentially hazardous chemical processes. How well do these milled circuits compare with chemically etched circuits? And how well do both approaches correlate with predictions from computer-aided-design (CAD) programs? How well can milled-circuit prototypes be replicated when compared to mass-produced circuits made by means of chemical etching?To compare the effectiveness of the two circuit-fabrication approaches, prototype microstrip bandpass filters were fabricated on 4350 20-mil-thick laminate material with 1-oz. copper (Cu) cladding from Rogers Corp. (Chandler, AZ). The bandpass filter is a four-pole Butterworth design with an eight-percent bandwidth of 400 MHz centered at 5.25 GHz (Fig. 2). It is suitable for National Information Infrastructure (NII) receivers (Rxs). With the growth of cordless telephones and wireless local-area networks (WLANs) in the 2.4-GHz band, the 5.2- and 5.7-GHz bands have emerged as strong candidates for short-range communications.

The M/FILTER and SuperStar linear analysis programs from Eagleware (Norcross, GA) were used for the circuit design and simulation tools. Half-wave folded transmission-line elements were chosen for this design because they did not require plated-through via holes to the ground plane. Synthesis data from M/FILTER (Fig. 3) predicted a center frequency of 5.2 GHz with passband insertion loss of less than 2 dB and a return loss of 30 dB.

Filters were fabricated with both processes (Fig. 4) and measurements were taken. The measurements showed the first trail-center frequency to be 250 MHz above the design-center frequency. This was greater than expected, but still close enough to allow a comparison of the two fabrication approaches. It was determined that the rejection characteristics were at a maximum for what could be expected, given the design constraints of size and insertion loss for this type of filter circuit. The unloaded quality factor (Q) of the printed resonators is less than 100.

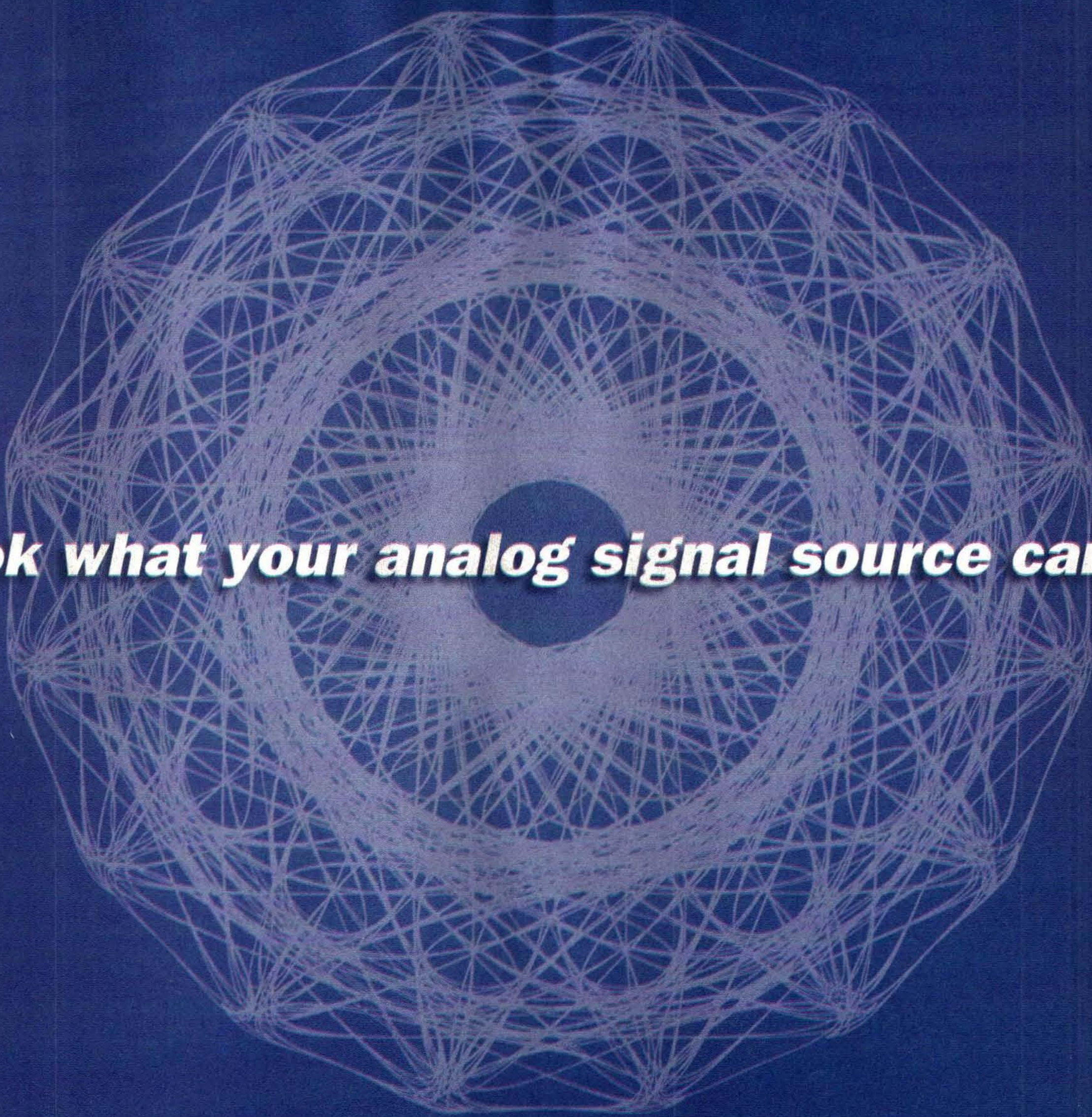
PRODUCING ONE'S OWN PROTOTYPE CIRCUITS (RATHER THAN SUBCONTRACTING THE WORK) THROUGH MECHANICAL MILLING HELPS PROTECT INTELLECTUAL PROPERTY (IP) WITHOUT RESORTING TO POTENTIALLY HAZARDOUS CHEMICAL PROCESSES.

Therefore, defining a 3-dB bandwidth of less than 10 percent would result in more than 3-dB insertion loss. Measurements of the two filter versions were made with a model 8720C automatic vector-network analyzer (VNA) from Agilent Technologies (Santa Rosa, CA)[Fig. 5].

The center frequencies in the milled circuits were closer to the computer-predicted value, while those for both etched circuits were a little higher. There were no superficial differences in the two types of fabrication, but microscopic examination revealed deviations from exact design dimensions of +0.5 to +1.0 mil in the mechanically milled filter and +2.0 to +5.0 mil in the chemically etched version.

The milled circuits (Fig. 6a) provided a more precise mechanical match to the original filter-design pattern because their traces were square and sharp, just as they were defined by electromagnetic (EM) CAD images. The etching process produced softer, more rounded edges (Fig. 6b). The filter bandwidth was within specified limits for all samples, but the insertion loss was greater than expected (approximately 2 dB was predicted by synthesis) in both cases: 5 dB for the milled filter and 3 dB for the chemically etched version.

DXF files provided by M-FILTER were used to prepare the plotter to manufacture the mechanically milled filter. The milled filters were produced in approximately 15 minutes



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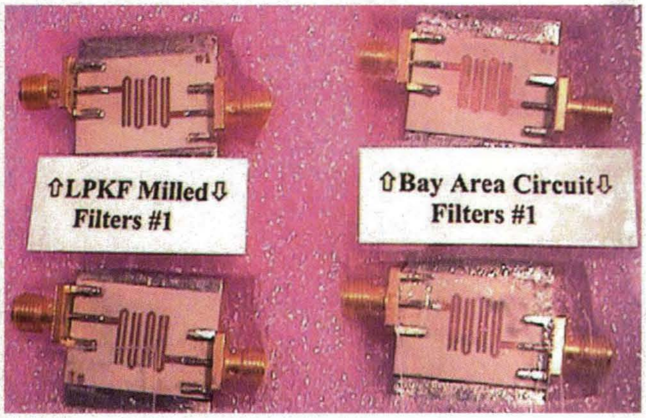
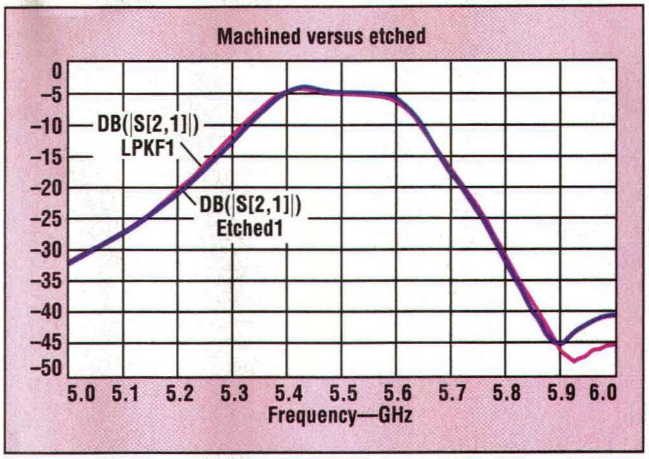


4. Using layout files from M/FILTER, mechanically milled and chemically etched filters were fabricated.



5. The measured performance of the machined versus the chemically etched filters was plotted with the help of a model 8720C VNA from Agilent Technologies.

using built-in software to generate the layout from the downloaded design files. The same design files were sent to a local vendor for the production of the etched samples, with a turnaround time of five days. Two samples were built with each process; all were made from the same Rogers 4350 laminate panel. The resulting test assemblies (Fig. 4) revealed that the mechanically milled circuits closely matched the electrical performance of the chemically etched filters, as well as the results predicted by EM software simulations. The simulations were performed with the aid of the Microwave Office software suite from Applied Wave Research (El Segundo, CA).

This project started with the question of whether prototype microwave filters produced by a mechanical process were equivalent to those fabricated by chemical etching. Since the goal of prototyping work is to arrive at a satisfactory design, the most important requirement is to have the circuits perform properly. Although the fabricated filters from both approaches were not identical in every respect, circuits from both processes did meet that requirement.

The significant difference between the two methods was turnaround time. Since most high-frequency designs are an iterative process, the ability to have a finished prototype in hand within minutes allows users to continue to improve a design in approximately real time. Waiting five days (including layout changes) for an etched board forces an engineer to set aside a design until the new circuit boards arrive from the outside service provider. The accumulated down time could become a substantial burden to a fast-track project.

The findings showed that milled circuits accurately reflect CAD-synthesis predictions and suggest that some minor corrections may be needed to account for those variations for replicating prototypes when using chemical etching for mass production. By considering the variations, a designer can be confident that a milled circuit will be as effective as a chemically etched version for the prototyping of new, higher-frequency devices. ••

Acknowledgments

The author would like to thank Bob Ritter, Chandos Rypinski, and Mike Hillbun for their substantial contributions to this project.

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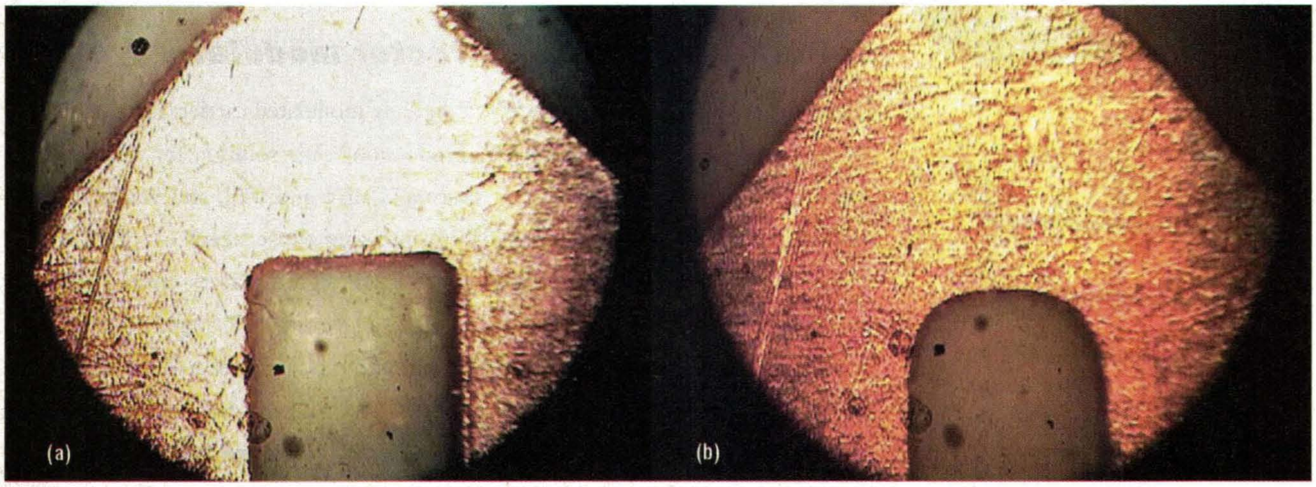
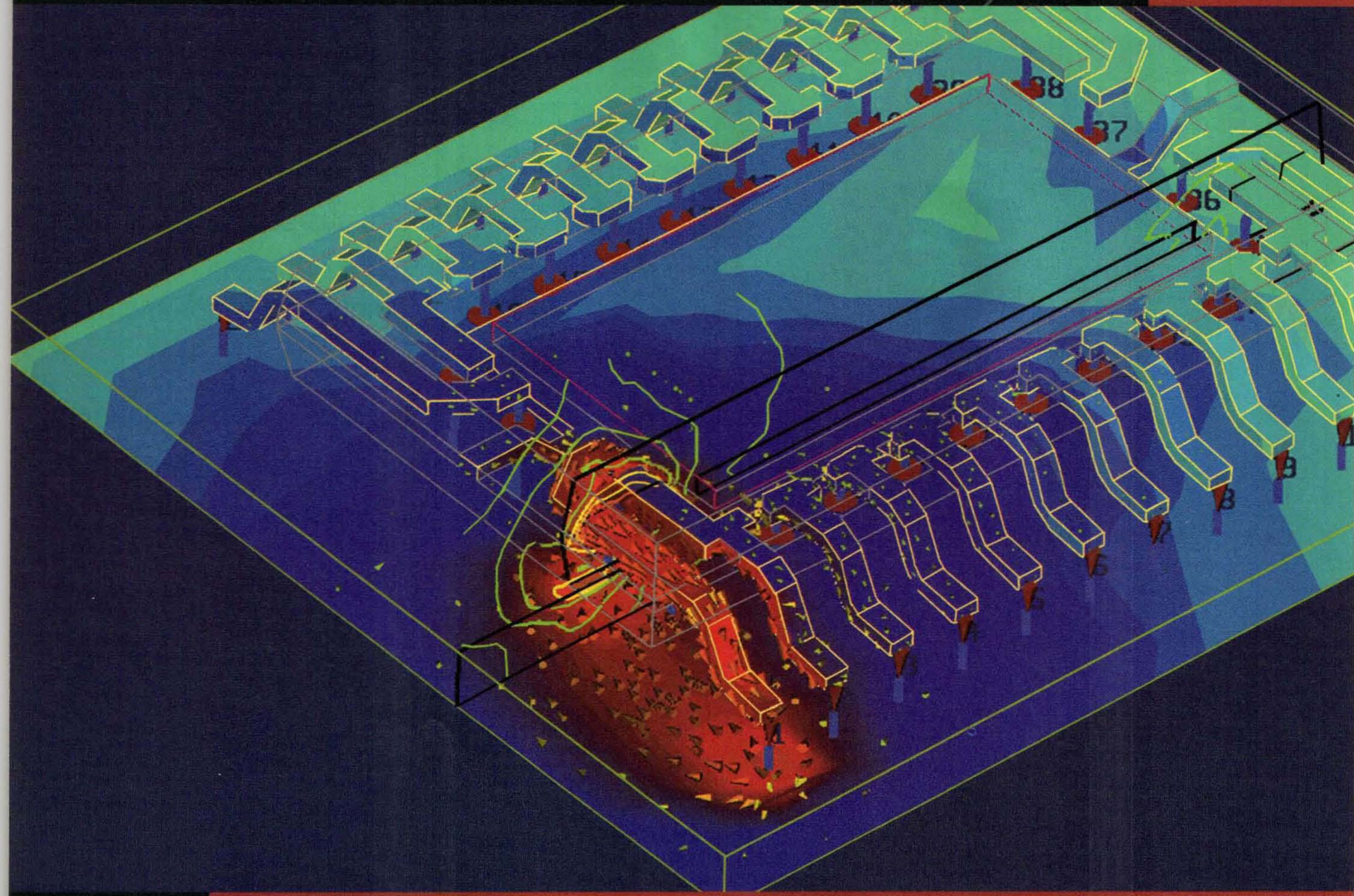


6. The mechanically milled filter circuits exhibited much sharper edges (a), in keeping with the design software requirements, compared to the rounded edges (b) of the chemically etched filters.



Comparing Infrared And Bluetooth Short-Range Solutions

Bluetooth and infrared short-range communications technologies can work well together when targeted to specific applications.

Evelyn Tay

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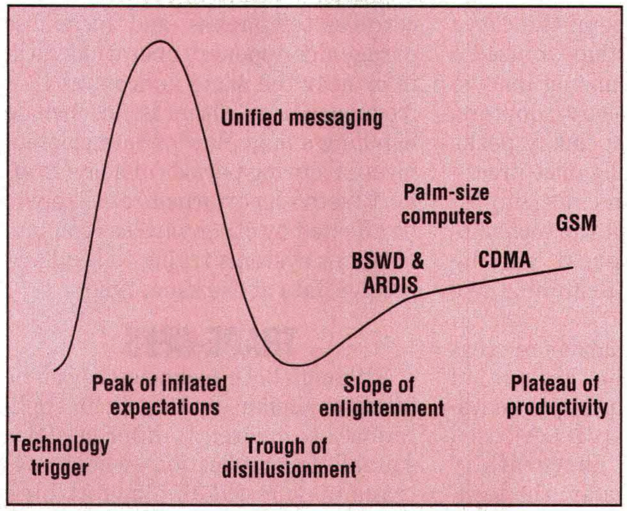
COMMUNICATIONS without cables has made computers with untethered peripheral devices possible. For short-range communications in the home or office, two technologies are currently available: Bluetooth and infrared (IR) devices. Both have their advantages and disadvantages, but both approaches can also provide complementary solutions for wireless data transfer between devices.

Today, the pervasive use of cableless data transfer has changed the way that people communicate. This trend is driven by the proliferation of mobile information appliances such as mobile telephones and handheld personal digital assistants (PDAs). In fact, further expansion of the market is expected with the growth of mobile Internet devices such as next-generation multimedia telephones, wireless-application-protocol (WAP) telephones, and smart telephones. According to Strategy Analytics, a high-technology market consulting firm, the global market for these devices will increase from a current annual level of $10 billion to an annual level of $73 billion by 2005. There is also a trend toward the convergence of capabilities in mobile communications and Internet access in order to provide mobile consumers with the latest information.

Although Bluetooth is often touted as a "replacement" for the IR devices of the Infrared Data Association (IrDA), the two technologies are actually quite complementary. The charter of IrDA for these low-power, IR devices is to provide low-cost, interoperable, market-adopted standards for communications between IrDA-enabled devices. Members of the IrDA include established multinational companies such as Agilent Technologies, Apple, Compaq, Casio, Ericsson,



Emerging technologies follow a predictable life cycle, with initial optimism boosting early market projections. (Artwork is reproduced with permission from the Andrew Seybold Group.)

Extended Systems, Hewlett-Packard Co., Microsoft, Motorola, Nokia, NTT DoCoMo, Palm Computing, Sharp, and Sony. The IrDA primary specifications include aligning devices point-to-point at an angle of 30 deg. and a distance of 0 to 1 m with data-transmission speeds between 9.6 kb/s to 4 Mb/s (with 16 Mb/s under development).

BLUETOOTH'S FUTURE

Bluetooth is an RF technology operating in the unlicensed 2.4-GHz industrial-scientific-medical (ISM) band. The Bluetooth operating standards were essentially established by a cellular-telephone developer, Ericsson. The firm later banded together to form the Bluetooth SIG with original members Nokia, IBM, Intel, and Toshiba, with the current expanded listing of Bluetooth SIG members numbering approximately 2000. Bluetooth specifications include omnidirectional voice and data transfer within a distance of 10 to 100 m at a maximum transfer rate of 1 Mb/s.

The tremendous attention that Bluetooth is currently receiving should not come as a surprise. According to the Andrew Seybold Group, technologies such as Bluetooth and IR typically follow a predictable life cycle, with tremendous optimism in the size of projected markets during the early adoption of the technologies (see figure). Bluetooth is on the upslope of the curve where the technology has aroused a high level of interest among innovators and early technology adopters. However, this interest finally peaks and dies down gradually once practical problems surface as implementers begin to adopt the technology. This dying interest will be revived only if the problems are solved.

Technologies that manage to cross the chasm have the chance of widespread adoption as implementation costs and risks drop. IrDA-based IR technology has successfully crossed the chasm. Today, the technology has solved its interoperability problems and gained widespread adoption in the marketplace. It is used in more than 150 million devices (as of 1999).

Bluetooth technology must address several issues before it can cross the same chasm. The current cost of implementing Bluetooth is much higher than the costs of implementing IR. In a 2000 article, Randy Giusto, the vice president for worldwide mobile research at the International Data Corp., states that the Bluetooth hardware cost alone is in the $20 to $25 range. IR, on the other hand, can be implemented at a cost of less than $5

Another issue is potential interference from other unlicensed RF devices. Unlike IR's performance, which is not affected by RF interference (RFI) or electromagnetic interference (EMI), Bluetooth must cope with interference from household appliances such as microwave ovens, cordless telephones, and automatic garage-door openers operating within or near the same unlicensed ISM frequency band. Other issues include handling a huge flow of data communication during periods of heavy traffic. Bluetooth performance is likely to be affected by other wireless communications systems trying to send and receive data at the same time.

ALTHOUGH BLUETOOTH IS OFTEN TOUTED AS A "REPLACEMENT" FOR IR DEVICES OF THE INFRARED DATA ASSOCIATION, THE TWO TECHNOLOGIES ARE ACTUALLY QUITE COMPLEMENTARY.

TRADE-OFFS

Although both technologies can be found in similar devices, their applications are inherently different. IR is suitable for applications where data transmission takes place at high speeds over a closely proximate line of sight paths. A good example would be data exchange between PDAs, notebook computers, and printers. On the other hand, omnidirectional Bluetooth devices may find it difficult to find each other even when they are in close proximity. Since Bluetooth transmissions are omnidirectional, carrying up to 100 m, they also lack the inherent line-of-sight security feature of IR. Files being exchanged will probably need to be encrypted and, possibly, password-protected.

Another application where IR is more suitable than Bluetooth is in the area of medical and health care. IR technology, unaffected by RFI, ensures that data transmission can be executed safely and effectively. Bluetooth will have potential problems with RFI and EMI, as well as other medical devices, causing disruptions and data errors.

Bluetooth is more appropriate for applications where line of sight is not desirable. For instance, users could synchronize their telephones without having to remain stationary. Bluetooth technology can also penetrate solid objects, allowing users to place telephones in their pockets while synchronizing.

The Bluetooth-IrDA SIG was established to identify, specify, and promote ways that both technologies can work together to provide additional benefits to the user. According to the organization's convener, Glade Diviney, IrDA can assist Bluetooth in two main areas. IrDA can assist Bluetooth to connect rapidly with a targeted device. Bluetooth, which operates omnidirectionally, may have difficulty trying to detect its intended target master or slave unit. IrDA and Bluetooth can also work together to provide connectivity over varying distances. While Bluetooth can reach its target of up to 100 m, its speed is slower. A device could use IrDA for rapid data exchange when it is placed close to its recipient. Subsequently, the device could change to Bluetooth when its targeted device is more than 10 m away.••

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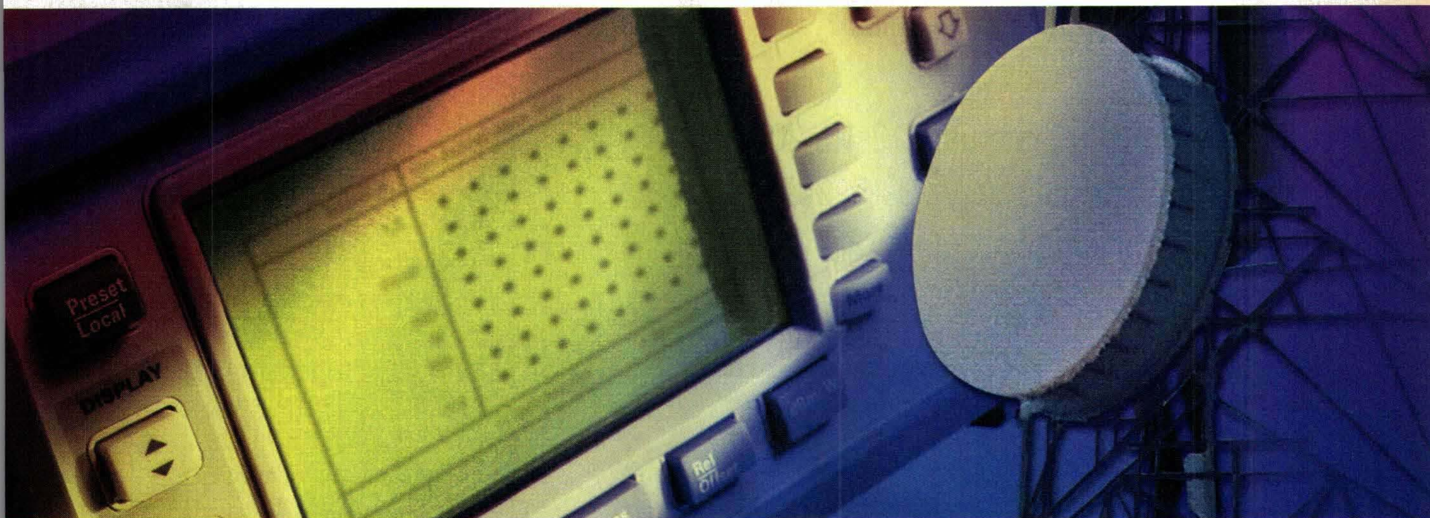

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AMID murmurs of hard and soft landings and reduced economic growth, the microwave industry continues to chug ahead. The only impediment to the growth of this market may be a shortage of the components needed to fuel it. At the very heart of wireless systems, the printed-circuit boards (PCBs) that lay the foundation for microwave circuits are being manufactured in unprecedented quantities. Like everywhere else in this market, cost and increased component density are driving PCB manufacturers to devise new solutions. However, increased interest in applications at higher frequencies is challenging the board fabricators with the most advanced capabilities to address the same high-volume manufacturing issues as their lower-frequency counterparts.

There are more than 750 companies in North America, Europe, and Asia capable of producing etched circuit patterns on microwave laminates. However, this list gets dramatically shorter when the challenge is to produce the complex circuits with the high operating frequencies required in some of the latest wireless handsets and prototype millimeter-wave circuits. Under these conditions, tolerances of a line or gap of approximately 12 μm are required, and even at lower frequencies, such precision is required for creating filters and tuned-circuit elements that do not require final tuning through trimming. At this point, the list of available suppliers drops by an order of magnitude or more, along with the magnitudes of product shipped.

Nevertheless, the mainstream business is keeping PCB manufacturers extraordinarily busy. "We are in a boom period from the wireless telecommunications industry," says Paul Koosmann, marketing manager at Modular Components (Forest Hill, MD, http://www.mcn-mmpc.com). "Business has been increasing faster in the last 3 years than in the last 10." Modular Components, is one of a comparatively few manufacturers that specializes in the fine line-etching and plated-through hole techniques required by millimeter-wave circuits in prototype as well as production quantities. The company handles line widths and gaps to 0.002 in. (0.005 cm) and tolerances to 0.0003 in. (0.0008 cm) on plated through-hole circuit boards. The company also can produce line widths and gaps with tolerances to 0.001 in. (0.003 cm) with deviations as low as 0.0002 in. (0.0005 cm) in prototype quantities.

Virtually every PCB manufacturer contacted by *Microwaves & RF* echoed Koosmann's confident sentiments. All attributed their steadily increasing business to the overwhelming demand for wireless products, and were willing to predict that this robust environment would continue at least through 2001 and perhaps beyond.

The key for many of these companies is how to accommodate this growth in a discipline that demands high skill levels. "Simply dumping a lot of money into a company and expecting good results is unrealistic," says Brit Andresen of Microwave Circuit Technology. "People are the key, and you just cannot train people off the street to do this kind of work." As a result, while Andreson's company is expanding, it is also staying close to its core capabilities, such as quick turnaround of prototypes. "We may have a small place in the market, but it seems to be a good place," Andreson continues.

At the other end of the spectrum, Polycircuits, Inc. (Bensenville, IL, http://www.mwav.com), continues to grow its business by expanding facilities in order to accommodate growth in the mainstream business for commercial wireless applications as well as higher-frequency circuits for emerging applications. The company produces double-sided boards in various finishes, multilayer circuits in up to 10 layers or five dielectric layers up to 100 mil thick, and mixed-dielectric circuits that incorporate different laminate materials in the same multilayer package.

This latter capability provides circuit designers with the flexibility to mix low-and-high-frequency digital

and microwave circuits to reduce space and manufacturing costs. The company has also developed a silicone (Si)-based conductive adhesive called Flexlink II for creating a thermally and electrically conductive bond between PCBs and metal carriers. This 0.0005-in. (0.0013 cm)-thick sheet of thermoset elastomer can withstand temperatures up to 260°C encountered in solder reflow. Since it is flexible, it is well-suited to use with polytetrafluoroethylene (PTFE), in which large dimensional changes occur. It has the same electrical properties as sweat-solder bonding.

Labtech Ltd. (Presteigne, Powys, United Kingdom, http://www.labtechcircuits.com) is another company that has diverse capabilities within the mainstream and higher-frequency markets. The company accommodates all aspects of the business, from quick-turnaround prototypes to high-tolerance high-frequency boards, as well as consultation services and seminars. The company also works with a large array of laminates, from Taconic TLC to RT/Duroid 5870 and 5880, aspect ratios to 8:1, drilling to 0.004 in. (0.010 cm), and plated-through holes with tolerances to ±0.002 in. (0.005 cm).

All of these companies are reliant on the ability of microwave laminate manufacturers such as Rogers, Taconic, and Arlon to deliver their materials in a timely manner. In the current environment of enormous demand, this has become exceedingly difficult, but manageable, with deliveries ranging range from a few days to two months or more.

HIGHER FREQUENCIES

While the bulk of microwave PCB business is concentrated on delivering product for wireless applications at 2 GHz or less, every manufacturer indicated that prototypes are increasingly requested for higher-frequency systems, including local multipoint distribution service (LMDS). This service, which operates between 27 and 30 GHz, has the potential to be the first large-scale commercial application for millimeter-wave products. This region of the spectrum—and beyond—has been viewed as ripe for development for many years, a feeling that has intensified as applications at lower frequencies have nearly saturated available spectrum.

However, the cost inherent in producing a system based on millimeter-wave components has made it the region of last resort for entrepreneurs looking for new applications and markets to develop. Virtually every microwave design challenge is magnified at these frequencies, where a full wavelength is only a fraction of an inch, and signal propagation can be impeded by precipitation. Consequently, only a situation in which lower frequencies are fully saturated could cause designers and investors to seriously consider producing cost-critical consumer products at these frequencies. However, this situation comes closer to reality every year.

LMDS is designed to challenge traditional wired solutions such as asymmetric digital subscriber line (ADSL) and cable modems for delivering high-speed Internet access, including voice, data, and video, to homes and businesses. Equipment mounted on the house or office communicates at millimeter wavelengths with a hub that serves subscribers in a small area, which in turn provides access to the public network. LMDS has the ability to deliver data rates of 45 Mb/s or more, which is in excess of what is currently achievable with either ADSL or cable.

Most analysts predict that the success or failure of LMDS depends on the ability of wired competitors to deliver the required bandwidth to businesses and consumers in a timely manner. As consumers trash their 56-kb/s dial-up modems and sign up for cable- or DSL-delivered high-speed access, they quickly recognize how truly useful the Internet can be, and they want more, including video on demand, which requires the highest possible data rates. LMDS can provide them with more, but not right now. However, since the proto-

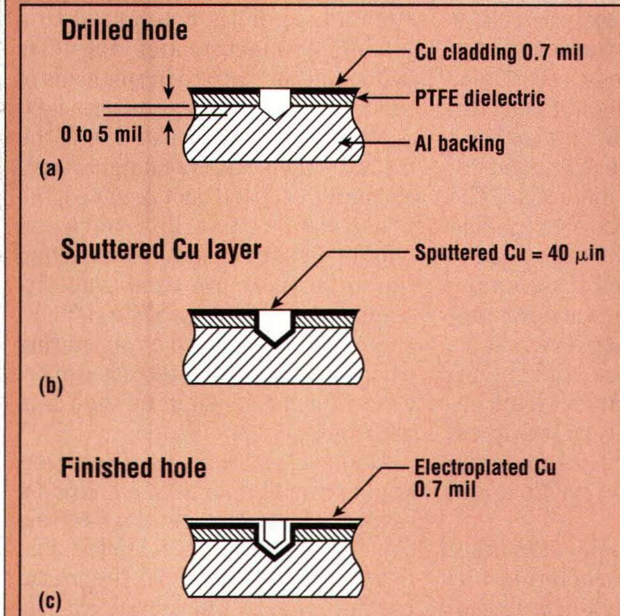
type requests from budding LMDS-system builders continue to increase, there is solid evidence that LMDS systems may begin to appear in large numbers in the near future.

"We have been working with LMDS customers for some time now," says Craig Sutton, CEO of Filtran Microcircuits, Inc. (Ottawa, Ontario, Canada, http://www.filtranmicro.com). "While the future of LMDS is uncertain, we are seeing increased demand for the types of precision circuits we build. Some of these people have indicated that quantities could increase substantially sooner than we might have expected." Unlike

AN OLD PROCESS RENEWED

While the machining properties of aluminum (Al) make it well-suited for three-dimensional (3D) shapes in high-power modules, it is electrochemically reactive and, thus, incompatible with most of the chemical processing used in printed-circuit manufacturing. Plated-through holes are an especially difficult challenge with this material. Of the various techniques employed to circumvent these problems, the "sputtered-blind-hole approach" developed by K. Ramachandran at Filtran Microcircuits, Inc. (Ottawa, Ontario, Canada) is one of the most effective.

Al requires extensive cleaning and surface treatments, principally zincating and anodizing, to form a barrier layer that isolates it from the plating chemicals. With zincating, the metal is dipped in a solution to form a thin coating of metallic zinc (Zn) on all the surfaces, followed by plating with electroless nickel (Ni) and cop-



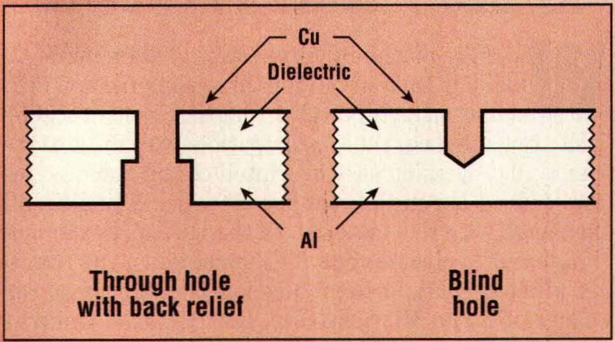
1. The process of creating a sputtered blind hole includes drilling, creating a sputtered Cu layer, and producing the finished hole.

some of its competitors, Filtran is actively ramping up for the greater volumes it believes are on the horizon from LMDS and other applications, such as adaptive-cruise-control systems, according to Sutton.

If the prototype requests of microwave PCB fabricators are indicative of future system development, then the next big wave of applications will be at millimeter-wave frequencies. LMDS, adaptive cruise control, battlefield and fire-control radar, and battlefield-communications systems all operate above 25 GHz, a region in which few components, PCBs included, are shipped today in large numbers at low cost. The next challenge for board fabricators will be to find ways to decrease cost and increase production quantities, while accommodating smaller and more complex circuits. ••

AN OLD PROCESS RENEWED

per (Cu). Anodization is achieved in an electrolytic bath by passing an electric current. Of the two, anodizing requires much tighter control of process parameters, but requires no third metal (zinc), which makes it more reliable for plated through-holes. In each case, plating coverage can be unreliable, reducing the aspect ratio of holes.

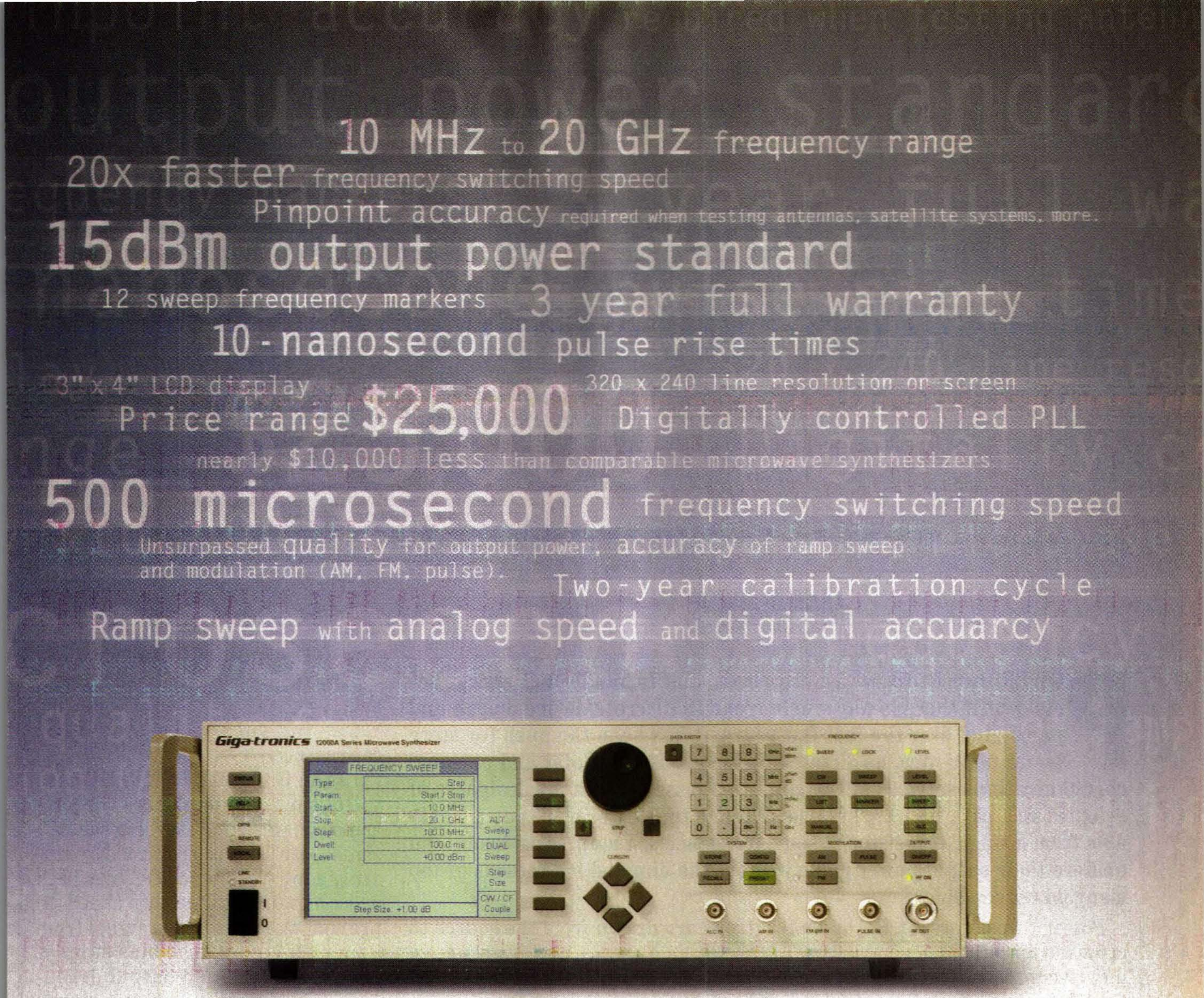


2. Blind holes eliminate the inherent limitation of sputtering—spraying around corners.

Another technique involves the insertion of tight-fitting brass, Ni, or stainless-steel pins wherever plated through holes are required. These pins are press-fitted and firmly anchored into holes passing through the top Cu foil and the polytetrafluoroethylene (PTFE) dielectric. The pin is soldered onto the surrounding Cu pad as part of the subsequent surface-mount assembly of components. Considerable mechanical precision is required to achieve reliable plated through holes, and the integrity of the mechanical joint between dissimilar materials is questionable under severe thermal shock, mechanical vibration, and corrosion.

Sputtered blind holes eliminate these drawbacks and can replace the chemical plated through-hole process on Al-backed substrates (Fig. 1). Sputtering has been used in the semiconductor industry for more than a decade, and employs vacuum metallization by a magnetron process. A molecular spray of metal inside a vacuum is generated when an inert gas, such as argon, is converted to a plasma by electric and magnetic fields, and bombards a target made of the metal or alloy. A substrate held against this spray is coated with the metal by molecular bonding. Since a chemical reaction is not involved, the process is compatible between widely dissimilar materials. When the surfaces are prepared properly, simultaneous coating of Cu, PTFE, and Al can be achieved by sputtering Cu inside and around the holes in order to form a continuous seed film without cracks or blisters.

Unlike chemical plating, however, sputtering cannot reach around a corner. This limitation can be overcome with shallow "blind holes" just deep enough to expose the metal backing (Fig. 2). They are much easier to drill than through holes and do not require the machining time required with conventional drilling.



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JACK BROWNE
Publisher/Editor

POWER transistors have always been built to handle high energy levels. But if there is any single trend in discrete-device design and packaging over the last few years, it has been the push for more power per device, as dictated by the needs of cellular and personal communications services (PCS) and, to a lesser extent, by designers of avionics and other military systems. With up to 1 kW or more power available from a single transistor at some frequencies, amplifier designers can achieve their output-power goals without the diminishing returns of too many power combiners.

Power transistors for RF and microwave applications generally fall into two materials types: silicon (Si) for RF power applications, and gallium arsenide (GaAs) for microwave power use. Recently, more exotic materials such as Si carbide (SiC), which is perhaps better known as the material behind blue light-emitting diodes (LEDs), and gallium nitride (GaN), have been evaluated for their capabilities in high-power RF/microwave transistors with encouraging results. And at least one company, Cree, Inc. (Durham, NC), has produced commercial products using SiC.

Traditionally, Si-bipolar transistors have been used for many high-power RF applications while GaAs metal-epitaxial-semiconductor field-effect transistors (MESFETs) have been used for power amplification at higher frequencies (through millimeter-wave frequencies).

In a GaAs MESFET, electrons are drawn from the source to the drain by a positive drain-source voltage. An input-signal voltage on the gate then modulates these majority electron carriers, producing voltage amplification. The maximum frequency of operation is limited by the length of the transistor's gate, denoted in micrometers (μm). The upper frequency limit or transition frequency (f_T, where the available gain reaches unity) of a GaAs MESFET can be found from:

$$f_T = g_m/2\pi C_{gs}$$

where:

f_T = the transition frequency,

g_m = the device transconductance, and

C_{gs} = the gate-to-source capacitance.

In the GaAs MESFET, DC bias voltage must be applied to the gate and drain without disturbing the RF-signal path. This is performed by creating a low-resistance DC bias path and a high-impedance path at RF, to prevent the microwave signal from being shorted by the bias supply. Similarly, input and output decoupling capacitors block DC from the input and output lines, allowing passage of the microwave signals.

While GaAs MESFETs are voltage driven, Si-bipolar transistors are current driven, with the base current modulating the collector current. The upper-frequency limit is determined primarily by the length of the transistor's base structure. This upper-frequency limit can be found from:

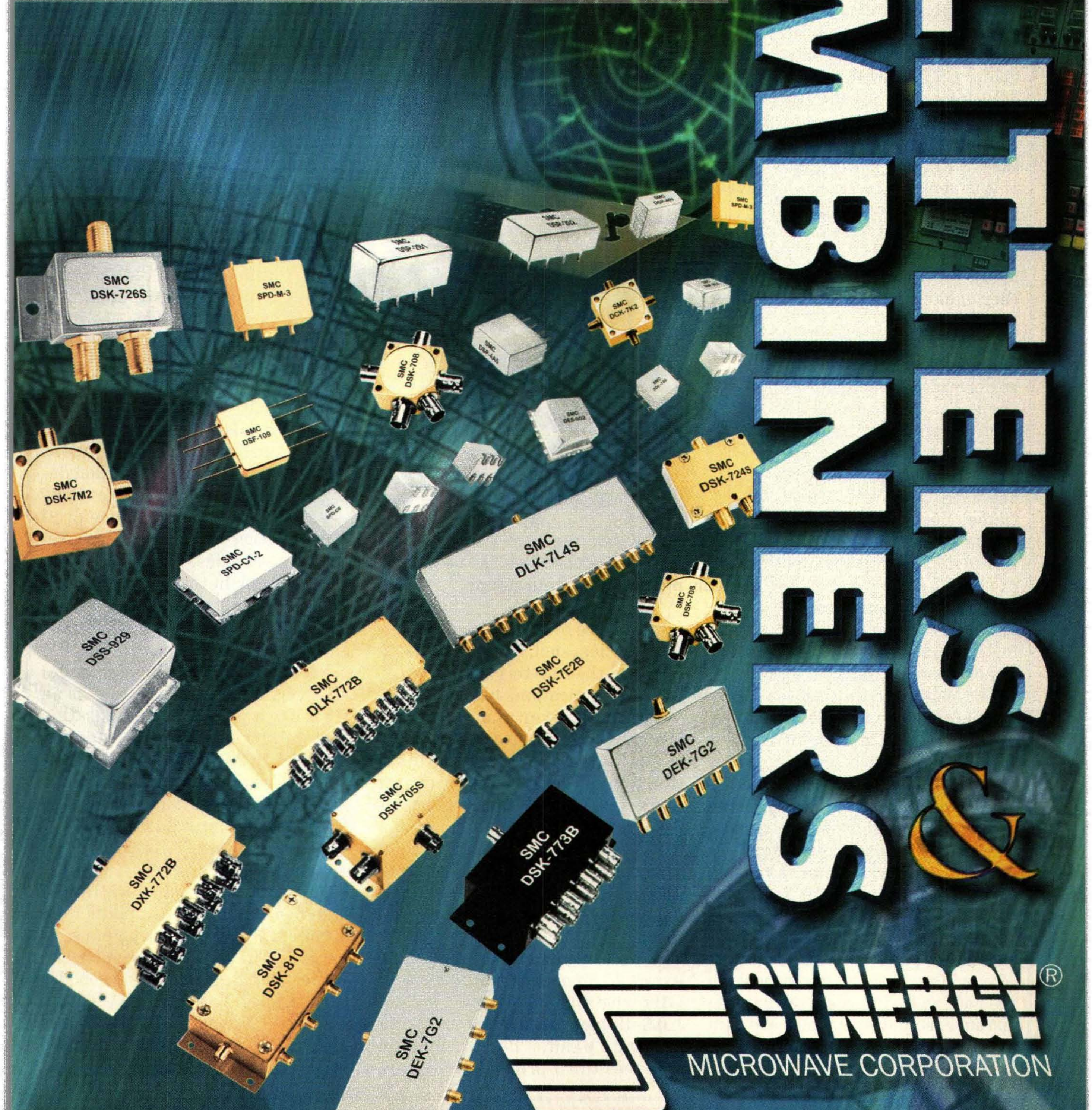
$$f_T = g_m/2\pi C_\pi$$

where:

C_π = the equivalent π capacitance.

Since the transconductance of a bipolar transistor is higher than that of a GaAs MESFET, the bipolar will typically provide higher gain than a GaAs FET at lower frequencies. But the larger capacitance in the bipolar transistor model serves to reduce the gain at

1. The MRF9045MR1 from Motorola is a moderately powered LDMOS transistor housed in a plastic package


SPLITTERS & COMBINERS
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SMC SPD-C1-2
SMC DLK-7L4S
SMC DSS-929
SMC DLK-772B
SMC DSK-7E2B
SMC DEK-7G2
SMC DSK-705S
SMC DSK-773B
SMC DXK-772B
SMC DSK-810

higher frequencies.

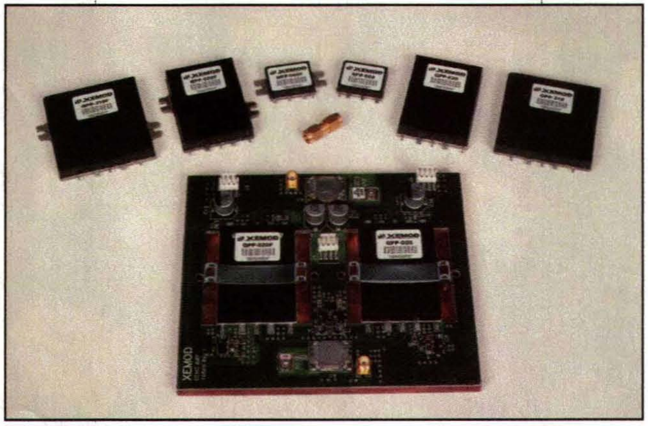
The charge carriers in the FET's channel can be either electrons or holes. If the carriers are holes, the device is known as a p-channel device. If the carriers are electrons, the device is known as an n-channel device. The structures of the two types of devices are similar, differentiated by a reverse in the flow of current and voltage. All FETs have a pinch-off voltage, which is the gate voltage at which the FET's channel turns off completely. In metal-oxide-semiconductor FETs (MOSFETs), the pinch-off voltage may be called the threshold voltage. If the pinch-off voltage is less than zero in an n-channel device, the FET is a depletion-mode device. If it is above zero, it is an enhancement-mode device. MESFETs are usually depletion-mode devices while MOSFETs can operate in either enhancement or depletion mode.

Parasitic circuit elements can affect high-frequency performance. In a MESFET, the gate resistance is one such parasitic element, with even a few Ω being significant. When combined with the gate-to-source capacitance, the gate resistance forms a resistance-capacitance (RC) filter at the input of the FET, reducing the gain at high frequencies. The gate resistance also generates thermal noise, and degrades the noise figure of the transistor.

The length and width of the gate is an important property of a FET. The shorter the gate, the better the high-frequency performance. The wider the gate, the higher the FET's transconductance and gate-to-source capacitance and the lower the source resistance. These parameters are roughly proportional or inversely proportional to the gate width. Increasing the gate width also increases the FET's maximum drain current, so power devices are invariably wide-gate devices. Increasing the gate width increases the gate resistance, so power devices are usually broken into a large number of smaller sections or cells connected in parallel. The gates of power devices may be several millimeters wide, and the structures of such large devices are often quite complex.

CHANGING WORLDS

GaAs MESFETs are still the dominant solid-state amplification device at frequencies above approximately 3 GHz. As unmatched devices, they are typically used in broadband applications such as amplifiers for military communications and electronic-warfare (EW) systems. In satellite-communications and terrestrial communications systems, however, internally impedance-matched transistors are more commonly used, in which the transistor has been prematched at the factory for optimum performance over a selected band of frequencies, such as the 3.7-to-4.2-GHz and the 14.0-to-14.5-GHz satellite-communications bands. Suppliers of high-power unmatched and internally impedance-matched microwave transistors include Fujitsu Compound Semiconductor (Santa Clara, CA) and Mitsubishi Electronics America (Sunnyvale, CA).



2. The QuikPAK modules from Xemod are impedance-matched amplifier circuits that can be readily combined to create high-power output stages.

Fujitsu, for example, offers a line of low-distortion internally matched power GaAs FETs ranging in frequency coverage from 2.5 to 2.7 GHz through 14.0 to 14.5 GHz and higher. The 25-W model FLM3742-25F, for example, operates from 3.7 to 4.2 GHz with 10.5-dB typical gain and ±0.6-dB gain flatness across the operating band. The transistor achieves third-order intermodulation distortion (IMD3) of –46 dBc. The device draws 4.8 A at +10 VDC.

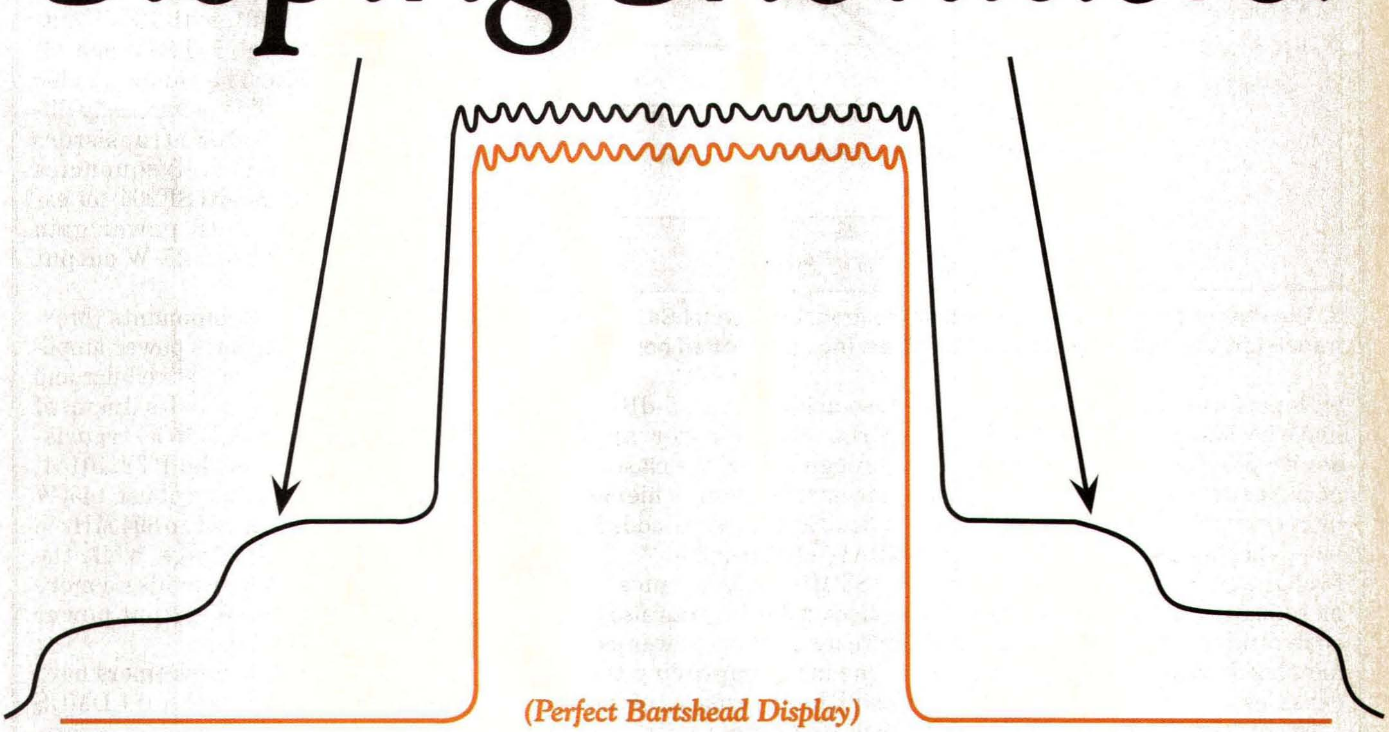
For the slightly higher-frequency satellite-communications band at 5.9 to 6.4 GHz, the firm also offers the FLM5964-25F, which also yields 25-W output power. It features 10-dB typical gain with ±0.6-dB gain flatness across the operating band, IMD3 of –46 dBc, and current draw of approximately 4.8 A at +10 VDC.

Mitsubishi's line of internally matched GaAs FETs includes devices through 14.5 GHz. The firm's model MGFC42V3742 is matched for use from 3.7 to 4.2 GHz and designed to deliver slightly more than 16-W output power with 10-dB gain. Over the frequency range of 3.6 to 4.2 GHz, the company's model MGFC45V3642A provides 32-W output power at 1-dB compression with power gain of 11 dB. Both transistors are designed for a power supply of +10 VDC.

Mitsubishi has also developed heterojunction GaAs FET devices capable of 160-W output power at S-band (compared to approximately 130-W output power in previous-generation devices), rivaling the output-power performance of Si power devices at those frequencies. The devices employ heterostructure-epitaxy technology, a new metalization system, and a T-shaped gate structure for improved power performance. The firm's model MGFS52V2122, for example, is an HFET in a push-pull configuration that delivers 160-W output power from 2.1 to 2.2 GHz for wideband-code-division-multiple-access (WCDMA) applications. It achieves 11-dB power gain at 2.12 GHz.

At frequencies below approximately 2.5 GHz, Si-bipolar transistors were once the darling of high-power solid-state amplifier designers. But, in recent years, Si MOSFET transistors have gained favor for their ruggedness and good IM performance. Variations on the basic FET structure have yielded diffused MOS (DMOS) and laterally diffused MOS (LDMOS) transistor types,

Danger. Sloping Shoulders.



Spectral growth can compromise the most precise CDMA technologies. Often, the difficulty lies in the transmission equipment, a potent source of distortion or the "shoulders" you see on your display. Now there are amplifiers that help you eliminate shoulders, keep the space between carriers narrow and optimize bandwidth.

AR "S" Series amplifiers are uniquely linear. Thanks to their exceptional design, CDMA signals get amplified but distortion stays low. You quickly pinpoint problems when testing your driver amp (or other transmission equipment) because you know the one place it's not coming from (the AR test amp).

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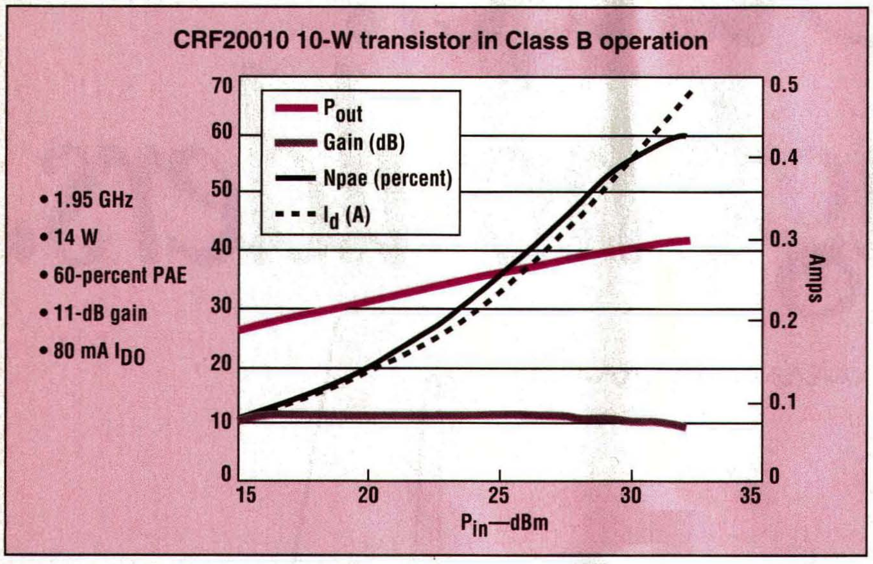

3. The output power, gain, and PAE of the first commercial SiC power transistor, the model CRF20010 from Cree, Inc., are plotted here.

with performance improvements in linearity and gain per millimeter of device periphery. And, while some manufacturers have employed aluminum (Al) metalization in their devices, the majority of RF FET manufacturers now use gold (Au) metalization in their devices to ensure minimal electromigration of metalized areas under high current densities.

In contrast to the vendor list for microwave GaAs MESFETs, the list of suppliers for Si RF FETs is well populated by companies such as Ericsson Microelectronics, GHz Technology, Mitsubishi, Motorola, Point Nine Technologies, Polyfet RF Devices, Stanford Microdevices, STMicroelectronics, UltraRF, and Xemod. One of the more popular device configurations for wireless applications below 2.5 GHz has been the LDMOS device. Capable of high power levels with low IMD (as required by the digital-modulation schemes used in many current- and next-generation wireless systems), these devices are typically used in base-station amplifiers designed for CDMA and WCDMA systems.

Motorola (Tempe, AZ), for example, has even developed a relatively high-power LDMOS transistor housed in a plastic package. The firm's model MRF9045MR1 (Fig. 1) is a +28-VDC device capable of 45-W peak-envelope power (PEP) while also achieving 18.5-dB gain at 945 MHz. Suitable for applications through 1 GHz, the plastic-packaged power transistor achieves IMD3 of −31 dBc and power-added efficiency (PAE) of 41 percent.

STMicroelectronics (Montgomeryville, PA) has also invested in research on high-power packaging as a means of improving the thermal and RF performance of their devices (see *Microwaves & RF*, April 1999, p. 131). By developing a patented self-aligned cobal-silicide process to reduce the polygate resistivity, the company has created lines of thermally stable, high-power devices with long projected operating lifetimes. The firm offers LDMOS devices for applications through 1 GHz ranging in power from the 6-W model PD57006 to the 170-W model SD7170.

Supplying high-power MOS devices since 1988, Point Nine Technologies (Newbury Park, CA) offers single-ended and push-pull Au-metalized, metal-gate MOSFETs for applications through 1 GHz. Model D1018, for example, is a 100-W push-pull device with 10-dB gain at 500 MHz. The firm also offers the C204 push-pull device, which is capable of 120-W output power through 800 MHz with 10-dB gain. In single-ended devices, the power leader is the model D1006, with 120-W output power and 16-dB power gain through 175 MHz.

Polyfet RF Devices (Camarillo, CA) offers a wide range of LDMOS power transistors for applications through 1 GHz. The lineup includes the single-ended model LZ402, with 125-W output power through 500 MHz and 12-dB power gain, as well as the push-pull (balanced configuration) model LR401, with 130-W output power through 500 MHz and 14-dB power gain. The company also manufactures a line of vertically diffused MOS (VDMOS) transistors that are usable at frequencies through 1 GHz. Model SK204, for example, offers 10-dB power gain through 1 GHz with 25-W output power.

Ericsson RF Components (Morgan Hill, CA) supports power-amplifier (PA) applications at cellular and PCS frequencies with its lineup of Au-metalized Si LDMOS transistors. The firm's model PTF 10161, for example, offers a robust 165-W output power from 869 to 894 MHz in a double-flange package, while the model PTF 10160 provides a more conservative 85-W output power from 860 to 960 MHz.

Several relative newcomers have been added to the ranks of LDMOS suppliers in the last several years, such as Stanford Microdevices, UltraRF, and Xemod. Stanford Microdevices, a "fabless" semiconductor manufacturer which also supplies much smaller-signal GaAs heterojunction-bipolar transistors (HBTs) and Si-germanium (SiGe) devices, offers a line of high-power Si LDMOS devices suitable for use from 800 to 2400 MHz at power levels to 120 W. The company's model SL-12010, for example, is designed for PCS CDMA applications from 1800 to 2000 MHz, where it is capable of generating 120-W output power with 10-dB gain. The IMD3 is a mere –30 dBc. Currently under development, the company's SL-16510 promises 165-W output power from 800 to 1000 MHz with 10-dB power gain and –30-dBc IMD.

UltraRF, until recently a part of amplifier manufacturer Spectrian (Sunnyvale, CA), is now an independent supplier of LDMOS transistors

(continued on p. 160)

Build a balanced 600-W power amplifier

Power-amplifier (PA) design requires the use of well-conceived impedance-matching networks around a reliable transistor or two. An application note from STMicroelectronics (Montgomeryville, PA), "Balanced UHF Amplifiers," provides a step-by-step guide to the design and construction of a 600-MHz pulsed amplifier for ultra-high-frequency (UHF) applications. The amplifier, which operates from 400 to 450 MHz, features impressive 0.5-dB amplitude flatness across the operating band when operating with a 20-μs pulse width at a duty factor of 15 percent. The amplifier is designed for a power supply of +40 VDC.

The application note begins with the selection of a transistor for this amplifier. Using the company's model SD1565, a rugged metal-oxide-semiconductor-field-effect transistor (MOSFET) with broad bandwidth, low thermal resistance, and better than 600-W output power with a 250-μs pulse width at 10-percent duty factor, the next step describes a technique to roll off the power at the band edges to a drop off of 0.5 dB. Since the 3-dB bandwidth has been determined as 400 to 450 MHz, the 0.5-dB rolloff bandwidth is determined by 3× the 3-dB bandwidth, or 3 × 150 MHz.

The note shows how to resonate out the inductance presented to the input of the transistor by adding a shunt capacitor with a value of 98 pF. Similarly, calculations show how to find the value of the capacitor needed at the transistor's output port. Transformations are then presented for achieving a balanced configuration with the transistor using a Chebyshev response with less than 0.04-dB ripple.

The note contains information on the basic amplifier layout, input- and output-matching networks, the calculation of required circuit-board (Teflon) thickness for optimum impedance matching, and the selection of transmission-line lengths for optimum amplifier performance.

The nine-page application note provides all of the details needed to construct a robust, 600-W pulsed PA at 425 MHz, working with the company's model SD1565 transistor. Copies of the note can be downloaded in text or probability-density-function (PDF) formats from the company's website. **STMicroelectronics, 141 Commerce Dr., Montgomeryville, PA 18936; (215) 361-6400, FAX: (215) 362-1293, Internet: http://www.st.com.**

For more information, visit www.mwrf.com

Improve Antenna Isolation With PCS Repeaters

Personal-communications-services (PCS) systems must provide wide area coverage at 1900 MHz, despite shadow-causing buildings and high-multipath conditions. For this reason, high-gain repeaters are often used as part of the PCS system to extend and improve radio coverage. An application note from WJ Communications (Palo Alto, CA), "Antenna Isolation With 1900-MHz PCS Band Repeaters," explains how to properly install donor and server antennas in repeater-based PCS systems.

To operate a PCS repeater system at maximum gain, the repeater system's antennas must be properly isolated from each other to minimize the opportunity for oscillation. The most common method of isolation is to mount the antennas some distance from each other. Since most installations are on a common tower, the antennas are normally mounted with vertical separation from one another. The server antenna is located high on the tower to achieve maximum coverage while the donor antenna is mounted lower on the tower within line of sight of the base-station antenna.

The note explains how to calculate the required amount of isolation for a particular amount of gain in the repeater. A simple formula is presented to translate the distance between repeater antennas into electrical isolation. An example is presented with repeater antennas from two different manufacturers in order to calculate the required isolation and then the required path loss between the two antennas. In estimating the isolation, it is found that a doubling of the distance between antennas at 1900 MHz adds an additional 6 dB of path loss.

The concise two-page note (Application Note 4) is available upon request from the company. It is also available as a free download from the company's website, along with another application note on achieving isolation with cellular-band repeaters. **WJ Communications, 3333 Hillview Ave., Palo Alto, CA 94304; (650) 813-2582, e-mail: Repeater-Support@wj.com, Internet: http://www.wj.com.**

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Preliminary Agenda

MONDAY, FEBRUARY 12

10:00 AM	The Demand for Wireless Applications
12:00 Noon	**Lunch Panel: Consumer Wireless Networks: Which Technology?**
1:30 PM	Infrastructure: Services, ASPs and Digital Bandwidth
3:30 PM	**New Appliances: Moving Beyond Free Phones**
6:00 PM	Dinner: Speaker to Be Announced

TUESDAY, FEBRUARY 13

8:00 AM	User Interface Challenges
10:00 AM	**Semiconductors**
12:00 Noon	Lunch: PDAs: Now and in the Future
1:30 PM	**Operating systems and software**

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COVER FEATURE

Broadband Synthesizer Trims Phase Noise Through 40 GHz

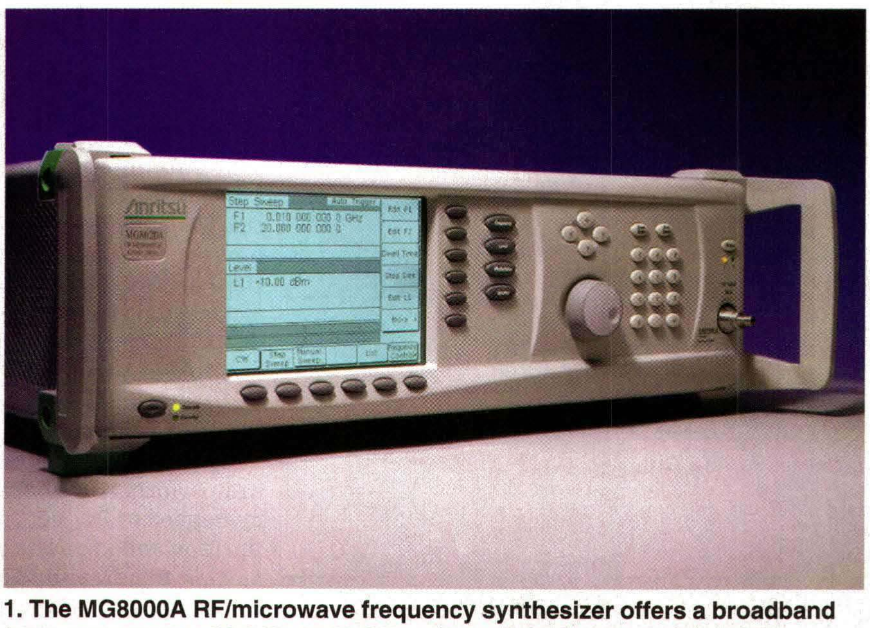
By adapting a compact modular construction with surface-mount components and integrating NCO technology, total RF and microwave coverage has been achieved.

JACK BROWNE
Publisher/Editor

MICROWAVE frequency synthesizers must provide superb spectral purity over extremely wide bandwidths when used in production and research and development (R&D) environments. The model MG8000A from Anritsu Co. (Morgan Hill, CA) does just that, using a digital downconverter (DDC) to achieve crystal-oscillator-like phase noise over a frequency range of 0.1 Hz to 40 GHz.

The MG8000A (Fig. 1) combines the bandwidths of separate RF and microwave signal generators with the spectral purity and frequency stability of a phase-locked source. The measurement-grade synthesizer achieves frequency resolution of 0.1 Hz over its full frequency range, with leveled output power that is adjustable in 0.01-dB steps from –120 to +17 dBm (when high-power Option 15 is installed).

The broadband-frequency synthesizer features outstanding phase-noise performance by combining a low-noise yttrium-iron-garnet (YIG) oscillator as the source of fundamental frequencies, a DDC module to generate low-noise signals from 10 MHz to 2.2 GHz, and a 48-b numerically controlled oscillator (NCO) to achieve the high-frequency resolution (Fig. 2). The basic architecture also includes several phase-locked loops (PLLs) for stability, along with an automatic-level-control (ALC) circuit that provides leveled output power over the wide amplitude range of the MG8000A. The DDC produces signals below 2.2 GHz by successive binary division of the YIG-source signals, thus avoiding nonharmonically-related spurious signals in the RF range below 2.2 GHz. Phase perturbations of the carrier are reduced with each successive frequency division.

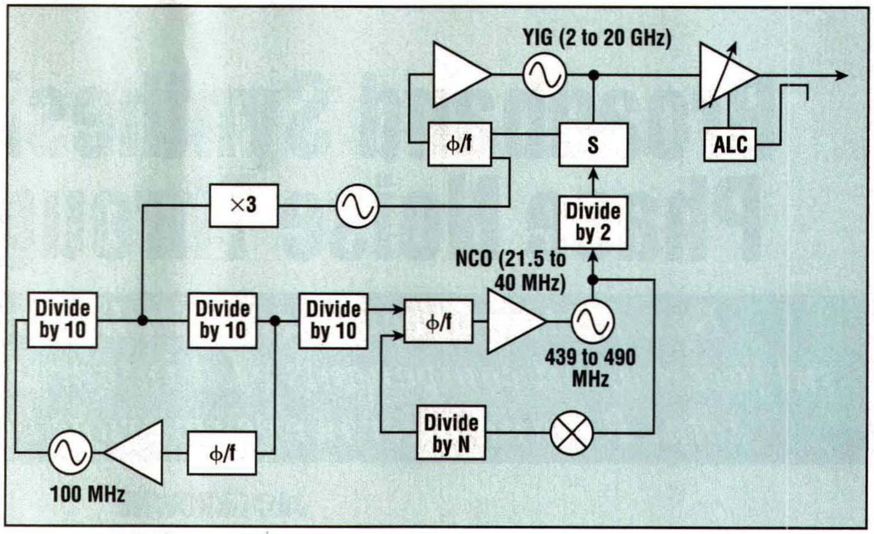


1. The MG8000A RF/microwave frequency synthesizer offers a broadband frequency range of 0.1 Hz to 40 GHz with low phase noise and various sweep modes.

When taking the cover off this rack-mount synthesizer, it quickly becomes apparent that, compared to the company's earlier synthesizer models, the MG8000A employs a lean layout with minimal clutter. Through the extensive use of surface-mount-technology (SMT) components, the 14 circuit boards and modules in the company's older frequency synthesizers have been reduced to only nine circuit boards and modules in the MG8000A. The use of SMT components, with their shortened lead lengths—which reduces package inductance—also helps to minimize noise sources in the MG8000A, and improve the overall spectral purity of the instrument.

The phase-noise performance of the MG8000A is summarized in the table and depicted graphically in Fig. 3. The single-sideband (SSB) phase noise of a standard MG8000A is -88-dBc/Hz offset at 1 kHz from a 6-GHz or lower carrier, rising to -75-dBc/Hz phase-noise offset at 1 kHz from carriers of 20 to 40 GHz. At 100 kHz from the carrier, the phase noise is -102 dBc/Hz when measured from a 6-GHz carrier and better than -100 dBc/Hz when measured from carriers to 20 GHz.

An option (Option 3) is available for improved phase noise performance. With this option installed, the phase noise is -101-dBc/Hz offset at 100 Hz from a 500-MHz carrier and -142-dBc/Hz offset at 1 MHz from the same carrier frequency. At a 6-GHz carrier frequency, the phase noise with this option is -107-dBc/Hz offset at 1 kHz from the carrier and -130-dBc/Hz offset at 1 MHz from the same carrier. The option provides for phase noise of typically -119-dBc/Hz offset at 1 MHz from a 40-GHz carrier. The MG8000A derives its frequency stability from an internal 10-MHz crystal reference oscillator with frequency stability of 10^{-7} and aging rate of 10^{-8}/year. It can also operate from an external 10-MHz reference with improved performance when better stability is needed.



2. A combination of technologies, including several PLLs and an NCO, are employed in the MG8000A to achieve a wide frequency range with minimum spurious, harmonics, and phase noise.

The MG8000A can be used as a continuous-wave (CW) source of single RF and microwave frequencies, or as a swept source, sweeping either frequency, power, or both. As a CW source, it features up to 20 independent markers, allowing an operator to set up to 20 independent CW frequencies. As a sweeper, sweep widths with the MG8000A can be set from as narrow as 1 kHz (0.1 Hz with option 11) to as wide as the full frequency range (40 GHz). The number of sweep steps can be adjusted from 1 to 10,000, with every frequency step in the range phase locked. The dwell time per step can be varied from 1 ms to 99 s. A number of sweep modes are available, including single-sweep mode and automatic sweep mode, along with a fixed-rate sweep mode in which the operator can adjust the total time of a sweep, including the lock time, from 20 ms to 99 s. The frequency-switching time is typically less than 15 ms + 1 ms/GHz step size, or less than 40 ms, whichever is less, to within 1 kHz of the final frequency.

The MG8000A also offers a list sweep mode, controlled through the front panel or remotely by general-packet interface bus (GPIB). It provides for up to four data tables with 2000 nonsequential frequency/power sets to be stored in memory and then addressed as a phase-locked step sweep. One table of 2000 points is stored in nonvolatile memory while the other tables are stored in volatile memory.

In addition, the MG8000A can perform a basic frequency-hopped or frequency-agile function under GPIB control. In this function, up to 3202 data points of power or frequency can be stored and recalled from nonvolatile

The MG8000A RF/microwave synthesizer at a glance

Frequency range	0.1 Hz to 40 GHz
Frequency resolution	0.01 Hz
Amplitude range Standard unit With high-power option	 -120 to +6 dBm -120 to +17 dBm
Amplitude resolution	0.01 dB
Amplitude accuracy	+ or -1.0 dB
Amplitude flatness	+ or -0.8 dB
SSB phase noise (at 6 GHz) Offset 100 Hz Offset 1 kHz Offset 10 kHz Offset 100 kHz	 -78 dBc Hz -88 dBc Hz -86 dBc Hz -102 dBc Hz
Spurious noise	< -40 dBc (typical)
Harmonics	< -40 dBc (typical)

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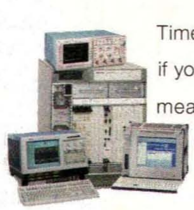

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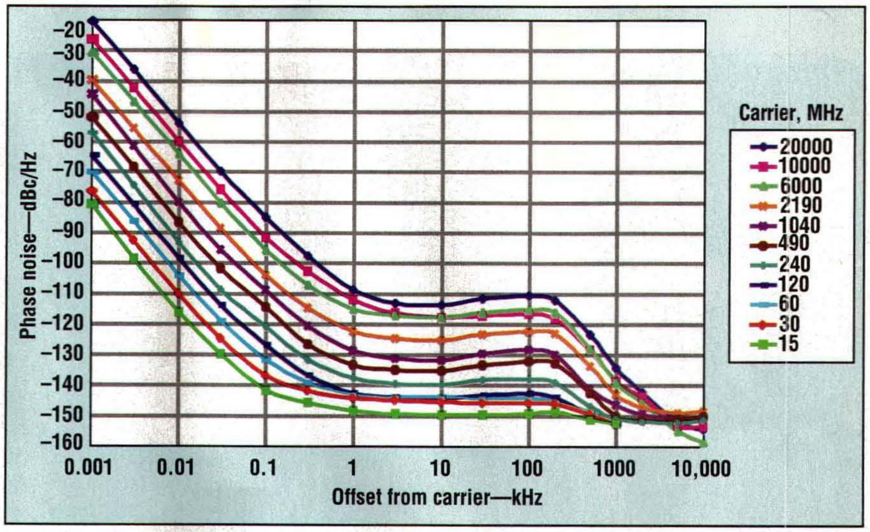

3. The typical SSB phase noise for the MG8000A is plotted for a variety of carrier frequencies from 15 MHz to 2 GHz and for frequency offsets from 1 Hz to 10 MHz.

memory. The switching time is less than 25 ms to be within 1 kHz of a final-hopped frequency.

PULSE MODULATION

The MS8000A incorporates pulse-modulation capability, with a pulse on/off ratio of better than 80 dB. The minimum leveled pulse width is less than 100 ns at carrier frequencies of 2 GHz or more and less than 1 ms for carrier frequencies below 2 GHz. The 10-to-90-percent rise/fall times are less than 15 ns (and typically less than 10 ns) for carrier frequencies from 10 MHz to 1 GHz and less than 10 ns (and typically less than 5 ns) for carrier frequencies from 1 to 40 GHz. Unleveled pulse-repetition frequencies (PRFs) are available from DC to 10 MHz while leveled PRFs are available from 100 Hz to 5 MHz. The level accuracy is ± 0.5 dB for pulse widths greater than 1 ms (at PRFs from 100 Hz to 1 MHz) and ± 1 dB for pulse widths less than 1 ms.

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InGaP HBTs Promise Long Operating Lifetimes

Ongoing studies show that GaAs heterojunction bipolar transistors with InGaP emitters offer reliable performance levels over long operating lifetimes.

Barry Lin
Senior Vice President for Manufacturing
EiC Corp., 45738 Northport Loop West, Fremont, CA 94538; (510) 979-8999, FAX: (510) 979-8902, e-mail: sales@eiccorp.com, Internet: http://www.eiccorp.com.

LONG life is a design goal for many semiconductor processes. But it is a goal that is apparently well met by a line of gallium-arsenide (GaAs) heterojunction bipolar transistors (HBTs) from EiC Corp. (Fremont, CA). Fabricated with indium-gallium-phosphide (InGaP) emitters, these GaAs integrated circuits (ICs) meet the needs of modern communications systems for smaller die sizes, better reliability, and better device-to-device repeatability.

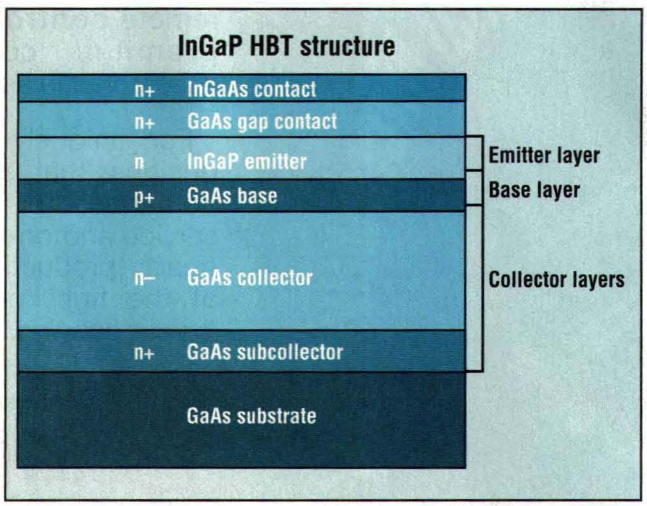
An example of the process is the EC-1019, a plastic-packaged broadband driver amplifier capable of 18.5-dB typical gain from DC to 3 GHz. With a noise figure of 5.5 dB, the amplifier yields output power of +19 dBm at 1-dB compression with an output third-order intercept point (IP3) of +34 dBm. The +5-VDC device is fabricated with the firm's proprietary InGaP HBT process.

Conventional aluminum-gallium-arsenide (AlGaAs) HBTs have been on the market for several years. In these devices, the p+ base region is generally fabricated by means of molecular-beam epitaxy (MBE) for beryllium (Be) doping. Unfortunately, the Be atom, being small, is unstable and diffuses rather quickly. Elevated junction temperatures and high junction-current densities accelerate this phenomenon, resulting in DC current-gain (beta) degradation. Also, the access-base surface near the emitter-base junction of an AlGaAs HBT is relatively unstable, requiring a careful surface passivation technique known as ledge passivation to reduce surface recombination effects. This surface effect leads to further beta degradation due to the increase in base current.

The first problem can be overcome by switching from Be to carbon-based dopant. Carbon is a larger and more stable atom. Unfortunately, with the high effusion-cell temperatures required to achieve carbon doping, this cannot be done with an MBE process, but requires metal-organic-chemical-vapor-deposition (MOCVD) epitaxy.

While the use of MOCVD carbon doping is a process improvement for GaAs HBTs, it does not cure the beta degradation due to the surface effect at the AlGaAs HBT base-access region, and the degradation is further accelerated by high current densities across the emitter region. This is especially true in applications where the current flow is continuous, such as occurs in optical communications systems.

Fortunately, InGaP-process technology can provide the benefits of HBTs without the shortcomings of an AlGaP process. The technology, long used in optical communications devices, employs an InGaP emitter in place of the less-reliable AlGaAs emitter. A cross-sectional view of an InGaP HBT device reveals an emitter-contact layer, emitter layer, base layer, and collector layer created atop GaAs subcollected and sub-



1. A cross-sectional view of an InGaP HBT reveals the InGaP emitter atop the GaAs base and GaAs collector regions.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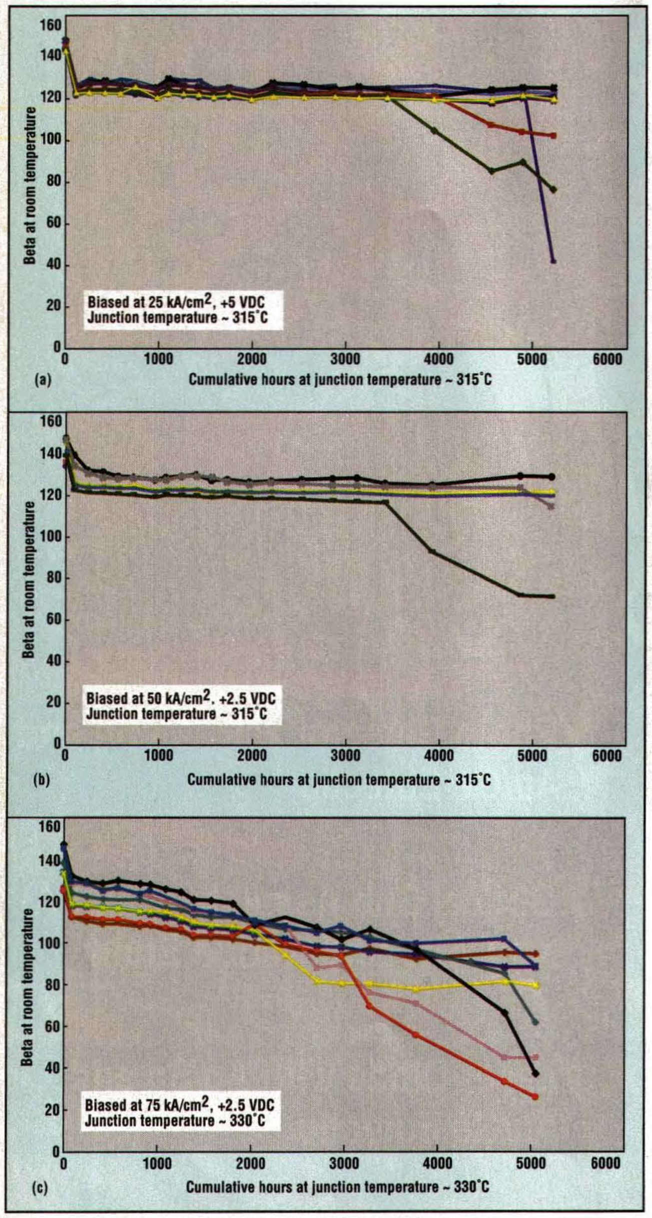
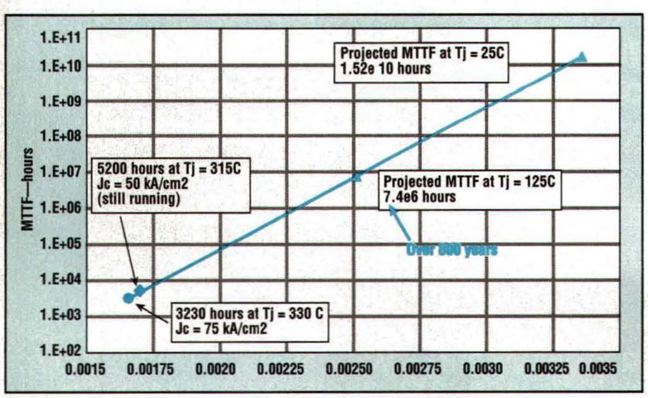


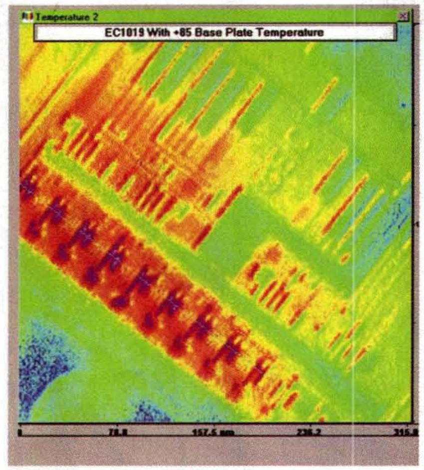
2. Lifetime tests were performed on the EC-1019 at three different profiles: (a) 25 kA/cm^2 of device area, (b) 50 kA/cm^2 of device area, and (c) 75 kA/cm^2 of device area.



3. The projected lifetime for InGaP HBTs at a normal junction temperature of +125°C is about 800 years.

strate layers (Fig. 1).

What are the reliability advantages of InGaP HBTs compared to AlGaAs HBT devices? In order to evaluate the mean time to failure (MTTF) of InGaP HBTs, stringent stress conditions were applied. Three sets of data were collected (Fig. 2a to c) based on measurements of devices mounted in ceramic packages. Devices were biased at 25 and 50 kA per cm^{2}of device area, at junction temperatures of 315°C, voltage levels of +2.5 and +5 VDC, with no failures reported in more than 6500 h of operation (Figs. 2a and b). Degradation finally occurred when the devices were operated at 75 kA/cm^2 at a junction temperature of +330°C. With 20-percent degradation of beta as the baseline measurement of failure, the MTTF for this level of stress (at 75 kV/cm^2) is approximately 3200 h.



4. Thermal plots of the EC-1019 show that the temperature gradients lie evenly distributed across the surface of the broadband amplifier

COOL TEMPERATURE = LONG LIFE

Based on these stress experiments, the projected MTTF is expected to be greater than 7.4 million h at a junction temperature of +125°C when operating at a current density of 75 kA/cm^2, using an activation energy of 0.8 eV (Fig. 3).

Under normal bias conditions, the InGaP HBTs maintain relatively cool junction temperatures, which supports the long MTTF. If proper thermal management is practiced, these devices can maintain low emitter-finger junction temperatures over extending operating periods, with negligible degradation in beta value.

Experiments performed on the EC-1019 have revealed even and controlled infrared (IR) temperature profiles (Fig. 4), with almost no "hotspots" across the full area of the device. This confirms the goal of maintaining a junction temperature of less than +125°C, achieved through the use of an InGaP process using MOCVD. **EiC Corp., 45738 Northport Loop West, Fremont, CA 94538; (510) 979-8999, FAX: (510) 979-8902, e-mail: sales@eiccorp.com, Internet: http://www.eiccorp.com.**

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Model, Analyze, And Simulate ΣΔ Fractional-N Frequency Synthesizers

Linear and nonlinear analysis techniques have been applied to the study of several fractional-N frequency-synthesizer architectures.

Part 2 of 2 parts

Yiping Fan

Senior Design Engineer

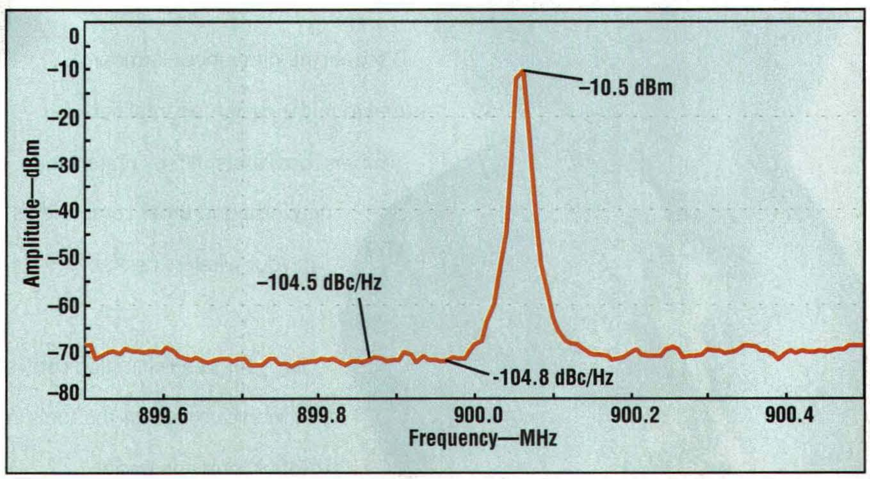
Part 1 of this two-part series concluded last month by indicating significant differences between ideal MASH11 and MASH111 fractional-N synthesizers with respect to idle-tone or spurious signal performance. However, a combination of shaping and loop-filter attenuation together with a fractional dividing ratio approaching 0.5 (for the MASH11) tends to reduce the spurious levels more than when the ratio lies closer to either 1 or 0. The simulations carried out to verify the analytical predictions, in fact, verify the location and amplitude of the spurious noise. In this second part, similar analyses will be performed on the MASH111 with the goal of illustrating the differences between the two architectures and also to demonstrate that good correlation exists between analysis and simulation.

A simulation was performed with the MASH111 structure (Fig. 11). There is no spur and the phase noise near the carrier frequency is flat. This is identical to the knee effect in Fig. 9. The phase noise is also slightly worse in the simulation than in the nonlinear analysis [in the nonlinear analysis, the folding effect beyond the half of the reference frequency is not counted due to the maximum frequency limitation of Fast-Fourier transform (FFT)]. A comparison of the analysis with simulation results shows good correlation (Table 2). As we indicated above, a fractional dividing ratio close to 0.5 will benefit the spurious performance for the MASH11 structure (see Fig. 12, where the same fractional dividing ratio is offset by 0.5.

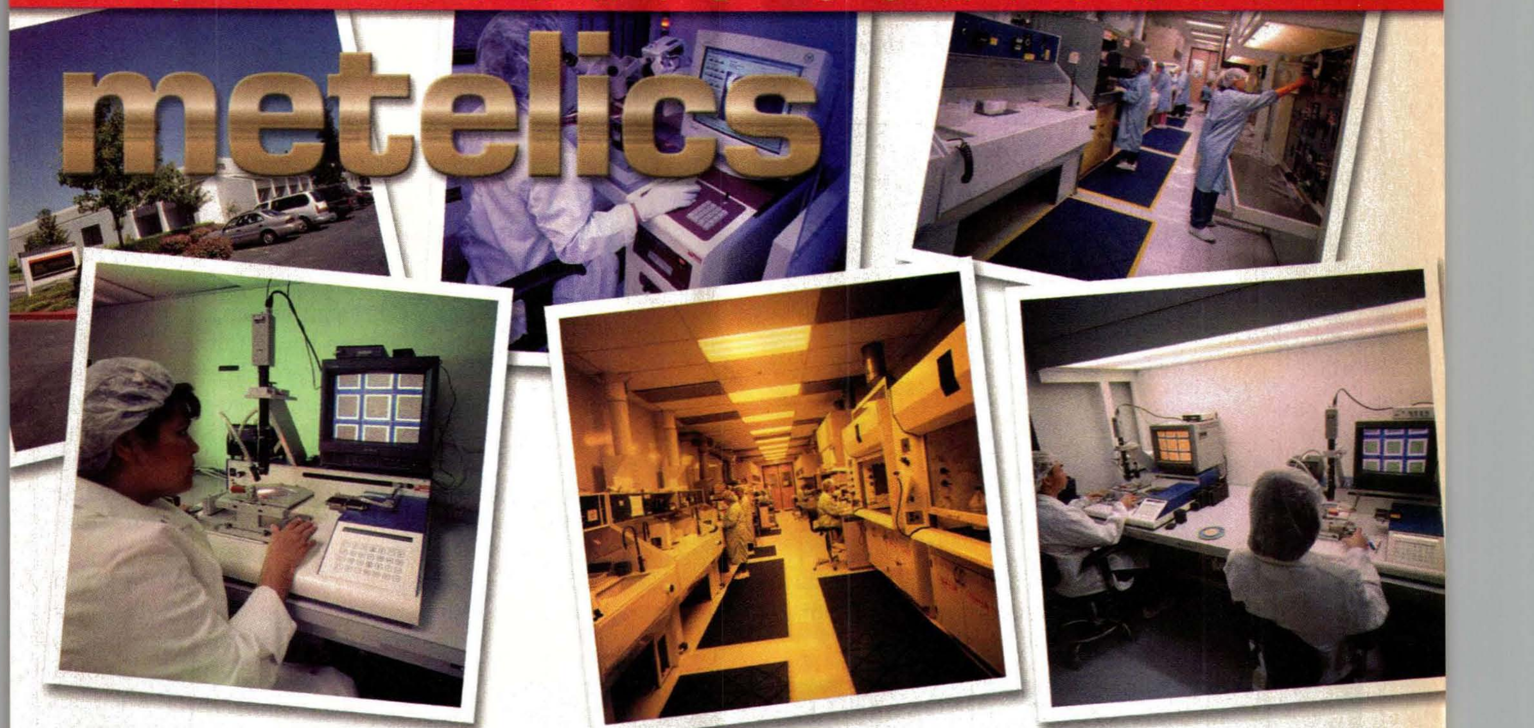
In a real synthesizer integrated circuit (IC), the circuits are less ideal and suffer from, among other things, nonlinearities and mismatch. A number of simulations were performed to show how a P and N charge-pump gain mismatch effects the spurious and noise performance. For example, Fig. 13 shows a cluster of plots when the charge-pump gain mismatch occurs for a MASH11 synthesizer. The carrier frequency is 900.060 MHz. There are five curves representing five mismatches of 10.0, 5.0, 2.5, 1.0, and 0 percent. The curves are coded by the gray levels. The lightest one has a 10-percent mismatch and the darkest one is for the 0-percent mismatch. There are not only spurs at 60 kHz but also at 120 and 180 kHz (Table 3). The increase in spurious level follows a slope of 20 dB/decade (for the mismatch). However, the close-in noise performance seems unaffected by the gain mismatch when the fractional division ratio is approximately 0 or 1.

MISMATCH TRADE-OFFS

Similar plots were made for a MASH111 synthesizer (Fig. 14), with the spurious levels for the different mismatches provided in Table 3. The



11. This plot shows the carrier spectrum of an ideal model for a MASH111 fractional-N frequency synthesizer.



Since its inception 21 years ago, Metelics has supplied microwave diodes for a vast array of commercial, military, and high-reliability applications. Today, Metelics is in the forefront of technology, with diodes that ride on commercial telecommunications satellites, the space shuttle, and a wide range of fixed and mobile wireless systems and test applications.

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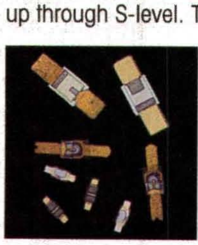
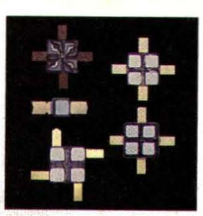


low Rd, very low capacitances, high uniformity, and are 100% tested and visually inspected. Diodes are available in rings, bridges, series T, antiparallel pairs, and singles. Surface mount packages are available in SOT 23, 0805, and epoxy-coated lead packages. Ceramic packages also available.

Pin Diode Chips and Beam Leads for Switch and Attenuator Applications

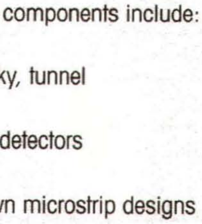


Metelics provides a wide selection of PIN diodes, SRDs, and varactors in chip, beam lead, and packaged configurations for RF and microwave requirements. Products range from very low-capacitance, high-speed switch devices to long-lifetime, high-power switch and attenuator types.

SRD (Step Recovery Diodes) and Tuning Varactors

Compared to packaged or chip devices, Metelics' silicon mesa beam lead step recovery diodes provide low-capacitance, very fast transition times, and low inductance, along with low parasitic capacitance. Tuning varactors are available from .5 to 20 pF (Cj4) in chips or packages.

RF/Microwave Components and Subassemblies



Metelics' years of experience in microwave diode beam lead and chip assembly are reflected in our enhanced performance, lower cost, and highly reliable state-of-the-art assemblies. Metelics' multi-device components include:

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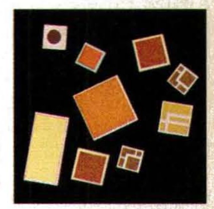
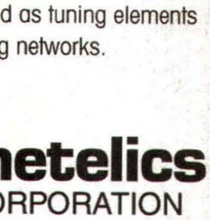


MIS (Metal-Insulating Layer-Silicon) chip capacitors have very high Q and small size for use in microwave hybrid circuit applications. Large bonding pads are supplied on most chips; the contact periphery is typically 2 mils from the edge, allowing wire or ribbon bonding near the edge for the lowest practical inductance. Beam lead MIS caps are also available.



Metelics' capacitors provide better performance than other types of ceramic capacitors, with low loss in supply decoupling circuits and GaAs FET transistor source bypass (providing more gain per stage). They can also be used as tuning elements in filters and matching networks.

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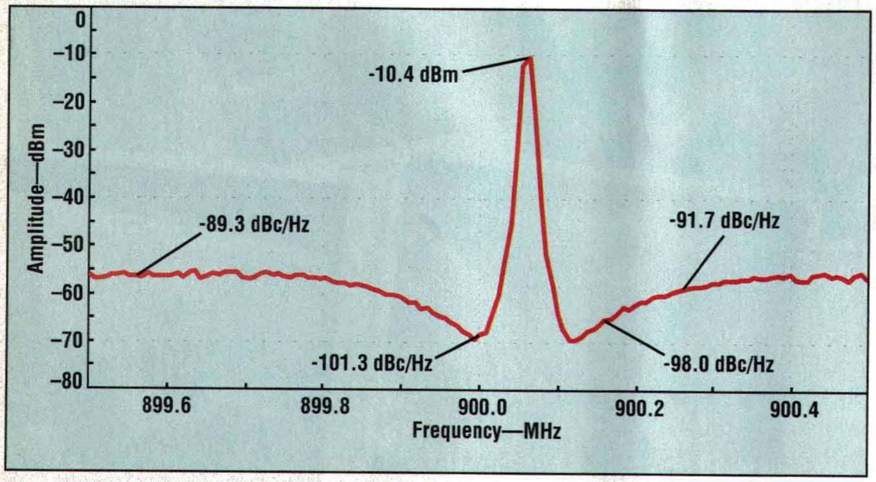

12. This plot shows the modified carrier spectrum of an ideal model for a MASH11 fractional-N frequency synthesizer.

Table 2: Comparing phase-noise performance levels of Figs. 9 and 11

Phase noises	Analysis	Simulation
at 100 kHz	−107.0 dBc/Hz	−104.8 dBc/Hz
at 200 kHz	−106.5 dBc/Hz	−104.5 dBc/Hz
at 500 kHz	−101.5 dBc/Hz	−100.8 dBc/Hz
at 1 MHz	−98.1 dBc/Hz	−97.4 dBc/Hz
at 2 MHz	−91.2 dBc/Hz	−92.8 dBc/Hz

Table 3: Evaluating MASH11 and MASH111 spurious levels

Mismatch (percent)	MASH11 at 60 kHz	MASH11 at 120 kHz	MASH111 at 60 kHz
10.0	−29.1 dBc	−39.6 dBc	−40.3 dBc
5.0	−35.0 dBc	−45.3 dBc	−46.1 dBc
2.5	−41.2 dBc	−49.6 dBc	−51.5 dBc
1.0	−48.8 dBc		
0.0	−52.9 dBc		

Table 4: MASH11 fractional spur and noise levels

Mismatch (percent)	Spur at 120 kHz	Noise at 60 kHz
10.0	−40.5 dBc	−94.4 dBc/Hz
5.0	−45.8 dBc	−98.6 dBc/Hz
2.5	−50.1 dBc	−101.3 dBc/Hz
1.0		−101.2 dBc/Hz
0.0		−101.1 dBc/Hz

MASH111 synthesizer is fractionally spurious when mismatch occurs. The spurious levels increase according to a rate of 20 dB/decade (mismatch). At the 10-, 5-, and 2.5-percent mismatches, the spurs at 60 kHz for the MASH111 unit are approximately 10 dB less than those for the MASH11 unit. However, the MASH111 pays a penalty in rising close-in noise levels of approximately 15 dB for the 10-percent mismatch and 10 dB for the 5-percent mismatch in order to achieve better spurious performance.

For the MASH11 structure, an ideal circuit will not generate spurs when the fractional-dividing ratio is approximately 0.5 (Fig. 12). When a mismatch occurs, this result is no more valid. The simulation results for the gain mismatch give rise to spurs that can be seen when there is at least a 2.5-percent mismatch. What is surprising here is that fractional spurious occurs not at 60 kHz, but at twice that fractional frequency, 120 kHz. The spurious levels are approximately 10 dB less when compared to those for a fractional-division ratio at approximately 0 or 1. The spurious slope is still 20 dB/decade (mismatch). There is another phenomenon not apparent in Fig. 13, namely, the fact that the close-in noise rises with the increase of the mismatch. This phenomenon suggests that for a MASH11 structure, when a nonlinearity exists within the loop, the spurious and close-in noise have a trade-off as the fractional dividing ratio varies from 0.0 to 0.5, or from 1.0 to 0.5. When a fractional ratio is approximately 0.0 or 1.0, the close-in noise is better and the spurious noise is worse, and vice versa as it nears 0.5. A summary of the spurious and noise levels is presented in Table 4.

It is critical to understand those phenomena related to the gain mismatch. Further analysis indicates that gain mismatch creates nonlinear effects. These nonlinear effects, in conjunction with $\Sigma\Delta$ phase noise, produce spurious signals and raise close-in phase noise levels. These nonlinearities are different from the VCO's nonlinearity, and the nonlinear effects due to gain mismatch can-

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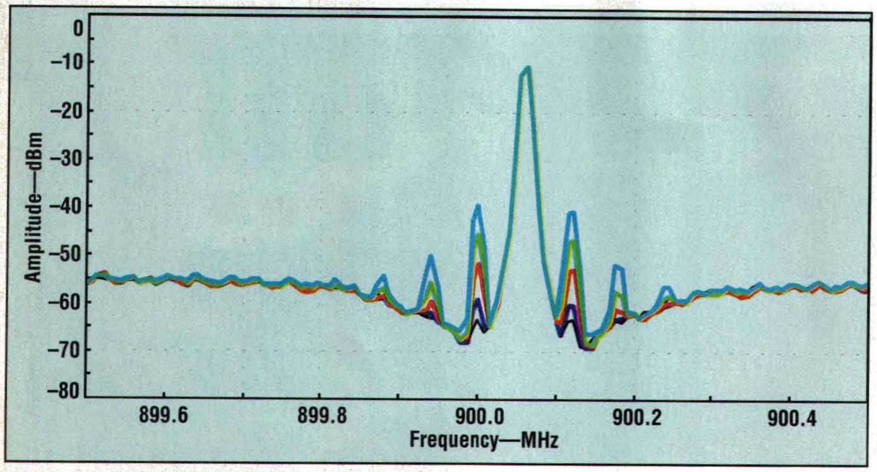
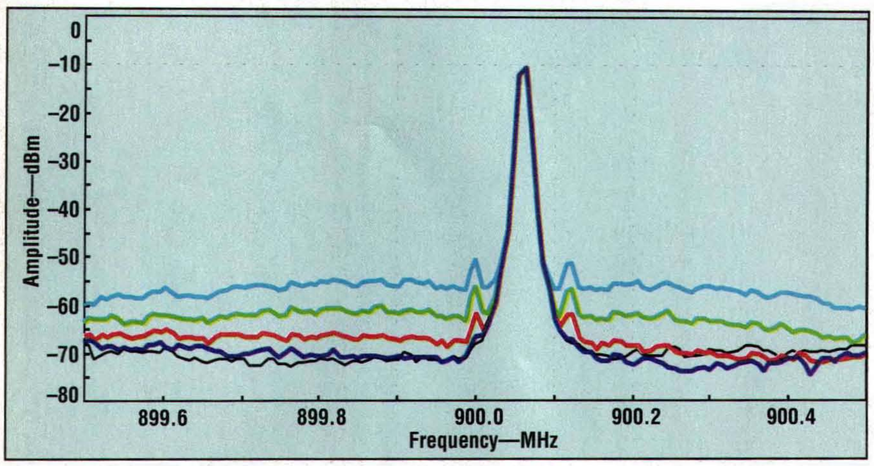
Table 5: Reviewing statistical $\Sigma\Delta$ phase noise

Phase error (cycles)	Fraction close to (MASH11)			Fraction close to (MASH11)		
	0	0.25	0.50	0	0.25	0.50
Maximum	0.9990	0.9980	0.9984	1.9883	1.9883	1.9883
Minimum	0.0000	0.0000	0.0000	0.0000	0.0000	0.0000
Average	0.0000	0.0000	0.0000	0.0000	0.0000	0.0000
Variance	0.1667	0.1670	0.1672	0.4995	0.4994	0.5000



13. This plot of the carrier spectrum for a MASH11 synthesizer includes the effects of charge-pump gain mismatch.



14. This plot of the carrier spectrum for a MASH111 synthesizer includes the effects of charge-pump gain mismatch.

Table 6: Comparing simulated and measured data

Spur/noise	Fraction close to 0 or 1		Fraction close to 0.5	
Spur	Simulation	Measurement	Simulation	Measurement
at 1 × fraction	−36 dBc	−36 dBc		
at 2 × fraction	−46 dBc	−48 dBc	−46 dBc	−47 dBc
Noise	Simulation	Measurement	Simulation	Measurement
Close in	N/A	N/A	−100 dBc/Hz	−97 dBc/Hz

not be reduced by the loop filter if the fractional frequency is within the loop bandwidth. In fact, this type of circuit imperfection can also be precisely modeled by a set of equations, although this will not be covered here.

The time domain may shed some light on the $\Sigma\Delta$ phase-noise properties. Table 5 provides some phase-error statistics for the MASH11 and MASH111 synthesizers at three fractional ratios near 0, 0.25, and 0.5. Note that the phase error is referenced at a carrier frequency. For the MASH11 synthesizer, the maximum phase error is approximately one carrier cycle, while for the MASH111 synthesizer the maximum phase error is approximately two carrier cycles. This is understandable, because the MASH111's output range is twice that of the MASH11 unit. The total noise power of the MASH111 synthesizer is approximately 4.8 dB higher than that of the MASH11 synthesizer. Differences in the time-domain statistics are small across the fractional range.

In summary, good correlation was made between simulations and actual measurements, using a model SA8528 $\Sigma\Delta$ fractional-N frequency synthesizer test chip from Philips Semiconductors [San Jose, CA] (Table 6). Careful attention was paid to the measurements in order to avoid influences other than the test circuit itself, to confirm that the correlation was genuine.

Acknowledgments

The author wishes to thank Rainer Gaethke and Ulrich Kohlschuetter from Philips Semiconductors, the main IC designers responsible for the SA8528 test IC, for providing measurement data and for many constructive discussions during the course of this study.

For Further Reading

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Startup Shaves Phase Noise From Microwave Sources

A diminutive supplier of oscillators and synthesizers seeks to vanquish spurious and phase noise from its lines of RF and microwave sources.

JACK BROWNE
Publisher/Editor

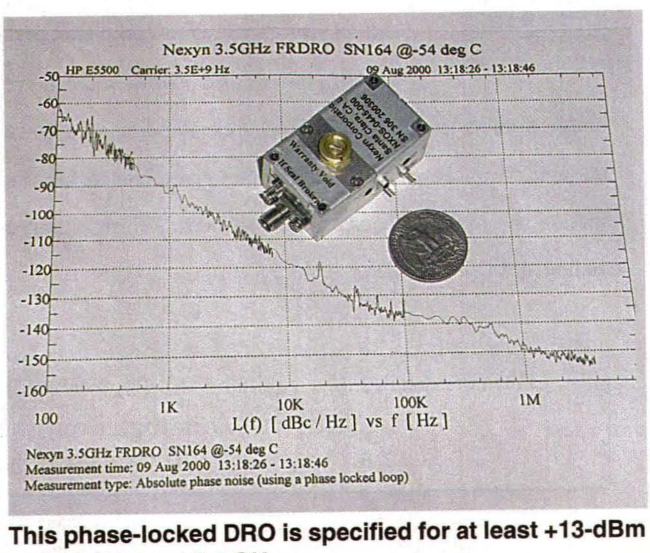
NOISE is the enemy of every receiver (Rx) designer. Fortunately, noise is under control at a tiny startup manufacturer, Nexyn Corp. (Sunnyvale, CA), where the company is producing free-running and phase-locked microwave sources with low phase noise, spurious, and harmonic content. To maintain quality control, the firm even manufactures its own oven-controlled crystal oscillators (OCXOs).

In spite of its small size, the firm offers a wide range of free-running and phase-locked dielectric-resonator oscillators (DROs) and voltage-controlled oscillators (VCOs). These sources are essentially frequency synthesizers and can be equipped with phase-lock alarms and other status indicators. An example of the firm's handiwork is the model NXOS-0350-01 phase-locked DRO, which is designed to provide at least +13-dBm output power at 3.5 GHz. Measurements reveal that the source (see figure) is actually capable of more than +15-dBm output power at 3.5 GHz. Although it is specified for output variations within ±1 dB, the DRO actually achieves output variations controlled within less than ±0.4 dB.

The NXOS-0350-01 offers a minimum mechanical tuning range of ±50 MHz. The pushing is typically 5 PPM/V, while frequency stability over temperature (–55 to +85°C) is specified as better than 3 PPM/°C. The source yields harmonic levels of less than –30 dBc and discrete spurious signals at less than –80 dBc. The phase noise at a 10-kHz offset from the carrier is typically–95 dBc/Hz, although measurements reveal performance as good as –116 dBc/Hz.

At higher frequencies, the firm offers the model NXPLOS-1900-01 a phase-locked DRO with better than +10-dBm output power at 19 GHz. It features less than ±1.2-dB output power variations with temperature (0 to +60°C). Designed for use with a 100-MHz reference oscillator, the 19-GHz source has measured spurious content of less than –75 dBc less than 200 MHz from the carrier and less than –65 dBc greater than 200 MHz from the carrier. Harmonics are specified at –25 dBc maximum, but measured at –35 dBc. The phase noise is typically –80 dBc/Hz offset at 100 Hz from the carrier, typically –97 dBc/Hz offset at 1 kHz from the carrier, typically –113 dBc/Hz offset at 100 kHz from the carrier, and typically –128 dBc/Hz offset at 1 MHz from the carrier.

Another example is model NXPLOS-I-1193-03-S, a phase-locked DRO with internal 10-MHz OCXO reference. The source is rated for at least +10-dBm output power at 11.925 GHz. With specified frequency accuracy of better than ±5 PPM (and measured performance of better than ±1 PPM), the source features better than –20-dBc harmonics and better than –60-dBc spurious content. The phase-noise offset at 1 kHz from the carrier is better than –70 dBc/Hz, improving to –110 dBc/Hz at an offset of 50 kHz. **Nexyn Corp., 678 Bend Dr., Sunnyvale, CA 94087; (408) 732-0793, FAX: (408) 730-0378, e-mail: sales@nexyn.com, Internet: http://www.nexyn.com.**



This phase-locked DRO is specified for at least +13-dBm output power at 3.5 GHz.

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(continued from p. 90)

It is obvious that for $\tau = NT_0 >> 1/\pi f_h$ this expression grows with N. Thus, the jitter grows with $\sqrt{N}$. This result also was obtained in:[5]

Flicker FM:

$$\Delta t^2_{RMS4} = \frac{T_0^2}{\pi^2} S_{\phi FFM} \int_0^{f_h} \frac{\sin^2(\pi f \tau)}{f^3} df = \frac{T_0^2}{\pi^2} S_{\phi FFM} \times (\pi\tau)^2 \int_0^{\pi f_h \tau} \frac{\sin^2(x)}{x^3} dx \qquad (24)$$

It is obvious that for $\tau = NT_0 >> 1/\pi f_h$ this expression grows with N^2. Thus, the jitter grows with N.

Random Walk FM:

$$\Delta t^2_{RMS_5} = \frac{T_0^2}{\pi^2} S_{\phi RWFM} \int_0^{f_h} \frac{\sin^2(\pi f \tau)}{f^4} df = \frac{T_0^2}{\pi^2} S_{\phi RWFM} \times (\pi\tau)^3 \int_0^{\pi f_h \tau} \frac{\sin^2(x)}{x^4} = dx \qquad (25)$$

It is obvious that for $\tau = NT_0 >> 1/\pi f_h$ this expression grows with N^3. Thus, the jitter grows with $N^{3/2}$.

The measured value of jitter is the square root of the sum of these "partial" jitter components:

$$\Delta t_{RMS} = \sqrt{\sum_{i=1}^{5} \Delta t^2_{RMS_i}} \qquad (26)$$

where:

Δt_{RMSi} are contributions of the power-law noise processes.

Part 2 of this article will appear next month. It will show how Eqs. 18 through 25 can be used to calculate jitter and how to evaluate the differences between measured and calculated jitter. ••

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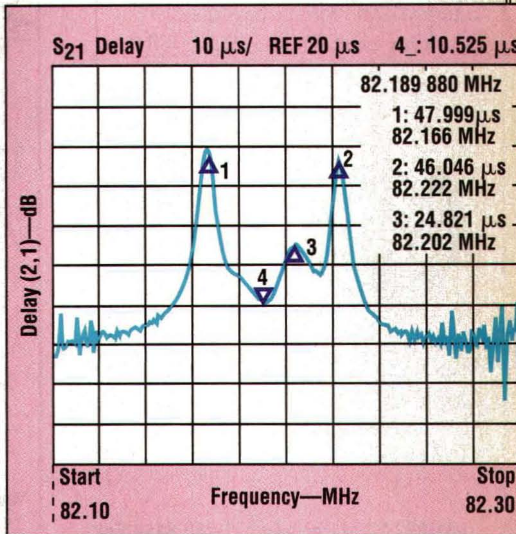
DESIGN FEATURE

(continued from p. 80)

MS8604A power meter.

The results of the measurements are given in Figs. 5 to 7. Figure 5 illustrates the S_{21} filter-response measurement. The value of S_{21} at the number 1 marker on the network analyzer is −3.1075 dB over the frequency range of 82.170 to 82.220 MHz. The comparable simulated response is shown in Fig. 4b—−3.332 dB, over the same 82.170-to-82.220-MHz range.



6. The group-delay response measured on the physical bandpass filter in Fig. 5 has the shape shown here. Compare with that of the simulation that is in Fig. 4a.

Figure 6 shows the measured group delay, which is to be compared with the simulated response of Fig. 4a. At the m1 marker in Fig. 4a, the simulated group delay is 16 µs (1.600E −5) at 82.19 MHz. The comparable measured group delay is 10.525 µs, also at 82.19 MHz. Figure 7 is a constellation diagram of the signal at the filter's output and its EVM. In the figure, the EVM value is 3.8 percent [3.78 percent root-mean-square (RMS) vector error in the figure]. Compared to the table in Fig. 4d, which reports the simulated EVM value as 3.7 percent, the experimental and simulated results closely coincide.

The method can be used for any filter with any center frequency.

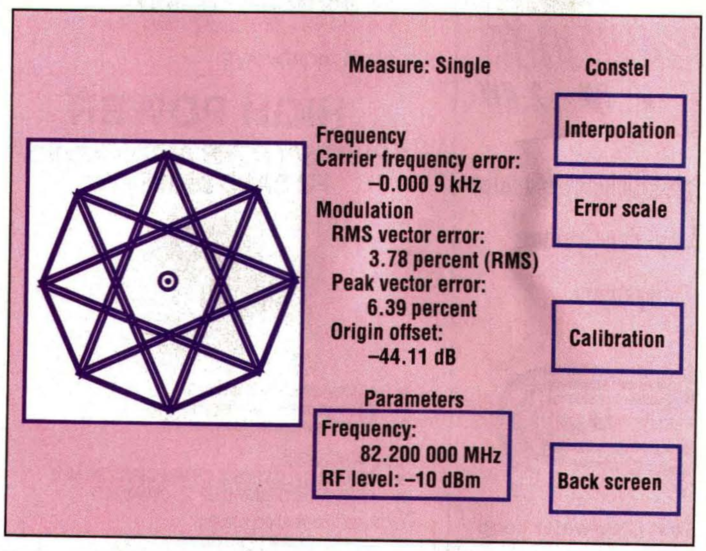
1. Set the frequency value "fc" at the variable and equations component "VAR1"on Fig. 1 to the central frequency of a filter (in MHz).

2. Obtain the S-parameters file of that filter and place it into the data directory of the ADS project.

3. Obtain an initial value of that filter's group delay either from measurements or from an RF bench simulation of the S-parameter file (obtained in step 2).

4. Sweep the group-delay value around the initial value (to obtain the minimum EVM).

The minimum EVM value is that caused by the filter's group delay. The ADS file of the subject matter of this article is available in a "zipped" form upon request. ••



7. A constellation diagram of the physical bandpass filter provides the EVM value shown by the RMS vector error figure—3.78 percent RMS. This value is virtually the same as that produced by the simulation (0.037 in Fig 4d).

Acknowledgment

I would like to thank my colleague W. Garner for suggesting the subject matter of this article.

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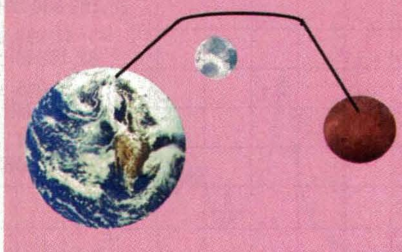

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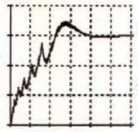

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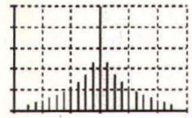

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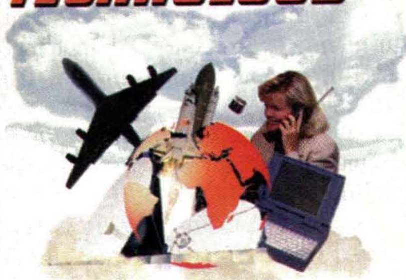

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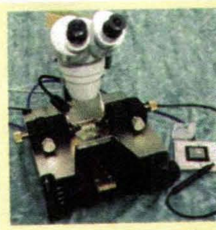

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(continued from p. 136)
or applications through 2 GHz. All of the transistors incorporate Au topside metal for high mean time to failure. Si-nitride passivation is standard for protection again surface contamination, while thermosonic wire bonding is used for reliability and repeatability from device to device. As an example of the technology, the company's model UPF18090 is designed for digital-communications-services (DCS) base stations operating from 1805 to 1880 MHz. Ideal for CDMA applications, the transistor produces approximately 90-W output power at +26 VDC with 12-dB power gain and 34-percent PAE. The company also markets model UPF18060, which is designed for the same frequency band, but with minimum output power of 60 W, 12.5-dB power gain, and 30-percent PAE.

Xemod (Santa Clara, CA) has taken the RF transistor to the next level by incorporating amplifier circuitry within its QuikPAC device packages. The power modules are impedance-matched Class AB amplifier stages designed for use in the output stages of high-power RF amplifiers, greatly simplifying the design of high-power amplifiers for cellular and PCS base stations (Fig. 2). The firm offers several amplifier/transistor modules at 120 W and more, including the model QPP-301 a Class AB power stage capable of 120-W output power from 2110 to 2170 MHz and 11.5-dB nominal gain. IMD3 is nominally –30 dBc, while the efficiency is nominally 31 percent.

The company also offers model QPP-017, a Class AB power stage with 200-W output power from 869 to 894 MHz. The QuikPAC amplifier module boasts 12.5-dB typical gain with 30-percent minimum PAE when driving 200-W PEP.

Although the increasing power levels in devices intended for cellular and PCS base stations is impressive, the real power levels are to be found in pulsed military applications, notably in avionics systems. GHz Technology (Santa Clara, CA) has long been a reliable supplier to military customers in need of high-power pulsed bipolar transistors, and offers devices exceeding 1 kW per transistor for short-duration pulses. A case-in-point is model ITC 1100, a transponder device for pulsed Interrogator systems at 1030 MHz, rated with a peak output power of 1 kW and 10-dB gain.

The company currently invests 15 percent of annual sales into research and development (R&D) for new devices and technologies. Model TAN 350 is a common-base bipolar transistor designed for pulsed systems from 960 to 1215 MHz. Aimed at air-to-ground and ground-to-air tactical-air-naviation (TACAN) applications, the transistor delivers 350-W peak output power, 70-dB power gain, 70-W maximum input power, +50-VDC, and 40-percent efficiency 10-μs pulse at 10-percent duty cycle.

The firm also offers model MDS 550L, developed for the extended-message MODE-S application (in Europe). The common-base bipolar device provides 550-W output power at +45 VDC and 1090 MHz. It is designed for maximum input levels of 90 W and will yield 7.8-dB power gain with 45-percent PAE when running from a +45-VDC supply. It is designed for an effective 128-μs pulse width at 2-percent duty cycle.

Model TCS 800 is a common-base bipolar transistor with 800-W output power at 1030 MHz for traffic-alert-collision-avoidance-system (TCAS) applications. Within transponder applications, the TCS can easily replace two 450-W transistors for reduced parts count and a savings in circuit-board space. The device, designed for maximum input-signal levels of 120 W, features 8-dB power gain, +45 VDC, 45-percent PAE, 32-μs pulse widths at 1-percent duty cycle. The devices are supplied in low-thermal-resistance packages for reduced junction temperatures and extended operating lifetime, and feature all-Au metalization.

Although a leading supplier to military customers, GHz Technology also provides devices for commercial customers. For cellular-communications applications, the firm's model 0809AB60 generates 60-W minimum output power from 864 to 894 MHz with 8.8-dB typical gain. The +26-VDC device offers 50-percent PAE. At PCS frequencies, the 1819CD60 provides 60-W output power from 1805 to 1880 MHz. The typical gain is 8 dB at +25 VDC. The transistor operates with 43-percent typical efficiency. For CDMA systems that require high linearity, the model 1920CD60 offers 60-W minimum PEP from 1930 to 1990 MHz with 8.5-dB gain and 43-percent typical efficiency.

If anything might change the status quo among the ranks of high-power transistor suppliers, it may be the fundamental acceptance of a new semiconductor device material, such as SiC. The material features a higher breakdown field that GaAs or Si, with performance comparable to that of gallium nitride (GaN). Due to this, transistors fabricated on the material support higher voltage ramps per length of periphery than GaAs or Si. Perhaps the most striking feature of SiC, however, it is impressive thermal conductivity, which is considerably higher than that of GaAs or Si. At 490 W/m-K, the thermal conductivity of SiC is, in fact, higher than that of Al (202 W/m-K) or even copper (Cu) [390 W/m-K]. With five times the power density of a GaAs FET, a SiC FET supports much smaller die size per watt of output power. The material supports transistors with operating voltages beyond +50 VDC, and exhibits high source and load impedances, enabling the creation of matching networks covering wide frequency ranges.

At present, the only supplier of RF power transistors based on SiC is Cree, Inc. (Durham, NC). The firm's model CRF20010 (Fig. 3) is a Class B FET capable of generating approximately 14 W output power at 1.95 GHz, with 11-dB power gain and 60-percent PAE. The linear device, with its excellent adjacent-channel-power-ratio (ACPR) performance in high-power amplifiers, is well suited for IS-95 CDMA base-station driver amplifier applications. And, although the output power of this device is relatively humble compared to some of the transistors mentioned earlier, the promise of SiC is great, and may lead to even smaller FETs at higher power levels. ••

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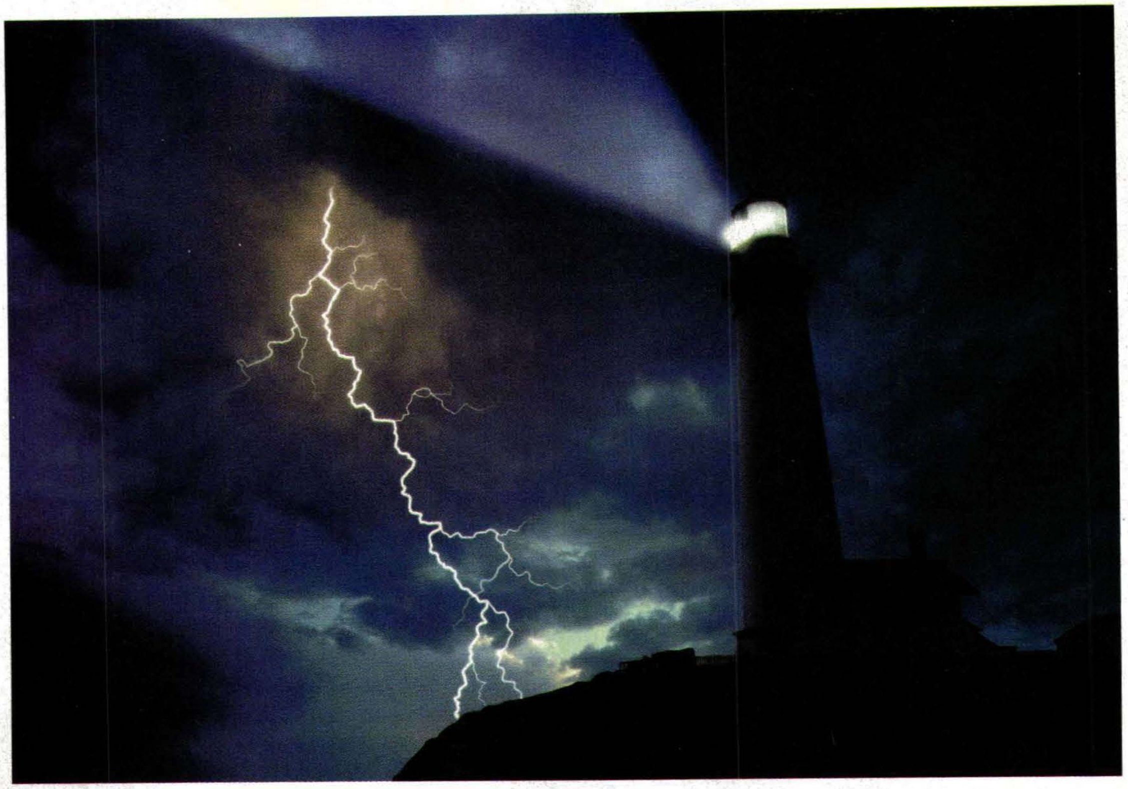

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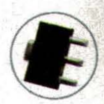

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A brochure features the company's surface-mount-device (SMD)/bonded-die hot gas-removal station and model 430 hot gas-jet add-on module. Specifications and application information are presented. **Semiconductor Equipment Corp.;** *(805) 529-2293, FAX: (805) 529-2193, Internet: http://www.semicorp.com.*

For more information, visit www.mwrf.com

*Domestic Edition only
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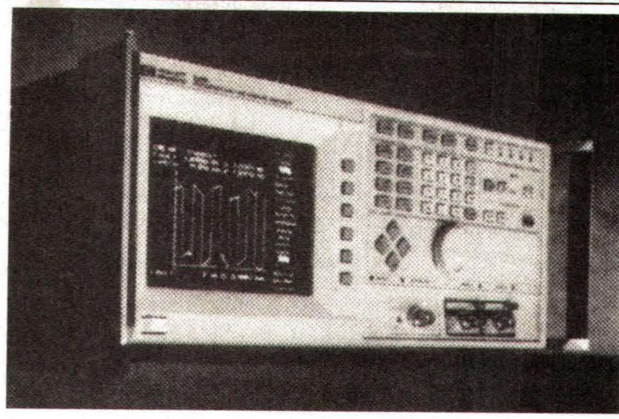

December 1987 marked the inauguration of this magazine's Top Products of the Year, with several noteworthy winners. For example, Hewlett-Packard Co. (now Agilent Technolgies) entered the new modulation domain of frequency-versus-time measurements with its HP 5371A frequency/time analyzer. And Wiltron (now Anritsu) made news with its first microwave vector network analyzer (VNA), the 360.

Microwaves & RF February Editorial Preview

Issue Theme: Fiber Optics

News

The speed of fiber-optic communications systems has jumped drastically in the last few years, from Synchronous Optical Network (SONET) rates at 622 Mb/s to current high-speed systems at 10 Gb/s. And device suppliers are already rolling out laser diodes and photodetectors for systems operating at 40 Gb/s and beyond. For an update on fiber-optic technology and how it impacts microwave-design strategies, do not miss this Special Report, prepared by Special Projects Editor Alan ("Pete") Conrad.

Design Features

The February issue of *Microwaves & RF* will blend the latest techniques in optical communications with traditional microwave-design methods. An author from Taiwan will investigate noise in power amplifiers (PAs) for Global System for Mobile Communications (GSM) networks. Another author from the old Soviet Union will look at the simulation of monopulse multipath effects at millimeter-wave frequencies.

Product Technology

February's Product Technology section will introduce a new line of high-performance RF/microwave integrated circuits (ICs) designed for low-cost, wireless applications. Additional articles will provide a close up on measurement solutions for Bluetooth testing and a line of high-speed fiber-optic components for systems operating past 10 Gb/s.